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tion Program and Summaries of Technical Papers appear in this issue.

IRE Standard on Designations for Electrical, Electronic, and Me-
ical Parts and their Symbols appears in this issue.

PROCEEDINGS OF THE I.R.E.

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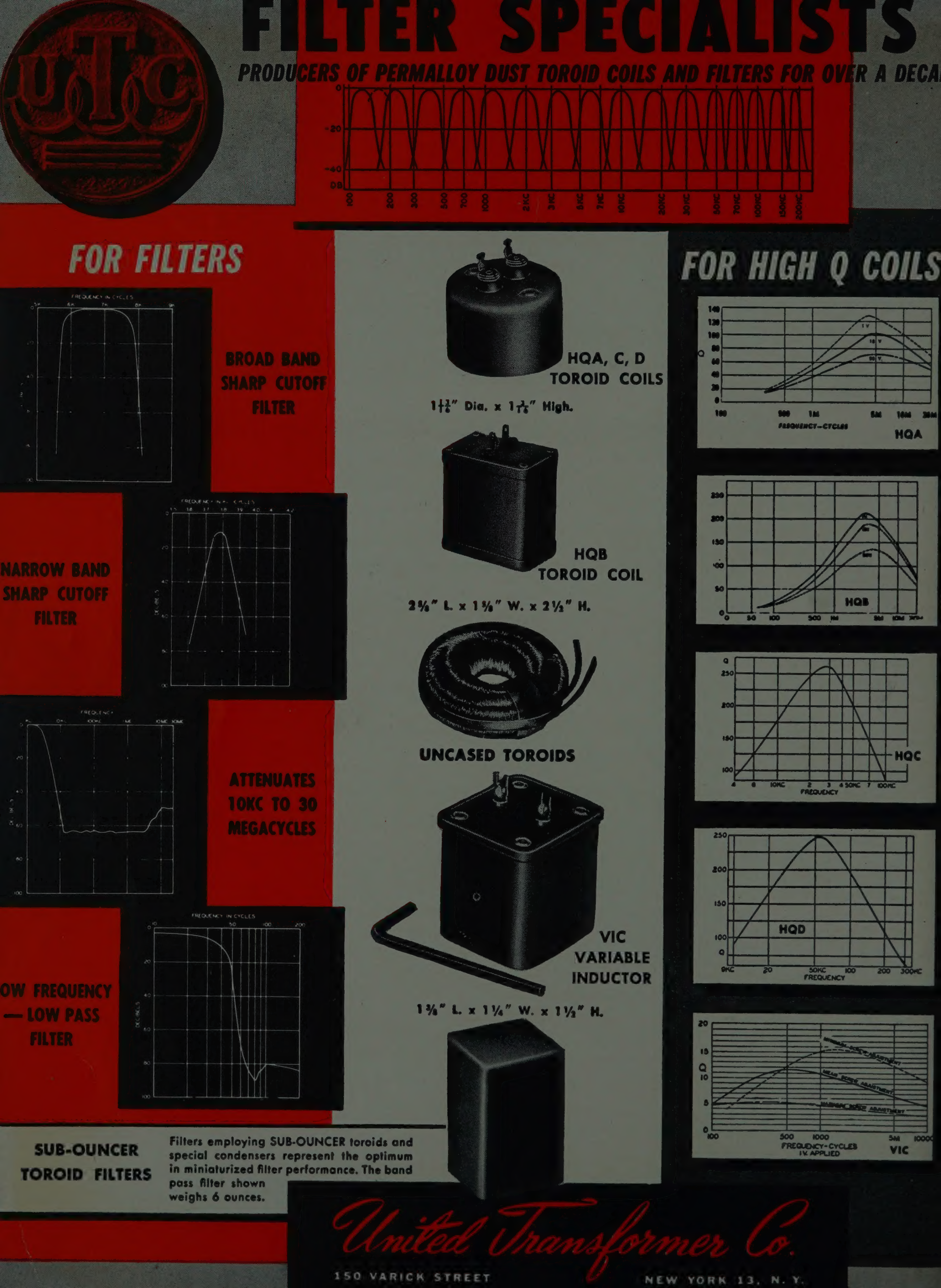

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PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

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Sir Robert A. Watson-Watt

VICE-PRESIDENT, 1950

Sir Robert Alexander Watson-Watt was born at Brechin in the County of Angus, Scotland, on April 13, 1892. He was educated at Brechin High School and at University College, Dundee, in the University of St. Andrews, graduating in 1912 with a special distinction in electrical engineering. He was invited immediately thereafter to become Assistant to the Professor of Natural Philosophy in University College, Dundee.

After teaching physics for a brief period in this capacity, he was appointed to the British Weather Service. In 1917 he became Meteorologist-In-Charge of the Branch Meteorological Office at the Royal Aircraft Establishment. A special purpose of this appointment was to enable him to investigate the possibilities of radio in the location of thunderstorms as a basis of thunderstorm warning to aviators.

He was later transferred from the Air Ministry to the Department of Scientific and Industrial Research. When the work of the Radio Research Station was merged with the radio program of the National Physical Laboratory in 1934, Sir Robert was appointed the first Superintendant of the Radio Department, National Physical Laboratory.

Sir Robert, who is considered England's foremost radar authority, was the leader of the earliest British work on radar, or radiolocation, as it was then called, and also was the first Director of Communications Development in 1938. In 1940 he became Scientific Advisor on Telecommunications in the Air Ministry and in the Ministry of Aircraft Production. He was Vice Controller of Communications Equipment in M.A.P. in 1942, and Deputy Chairman of the Radio Board of the War Cabinet, under the chairmanship of Sir Stafford Cripps in 1943. After the war he retained his appointment as part-time Scientific Advisor in Air Ministry, adding to it corresponding appointments in Ministry of Supply, Ministry of Transport, and Ministry of Civil Aviation. In 1947 he founded the scientific advisory and consulting engineering practice of Sir Robert Watson-Watt and Partners, Ltd.

His honors include his knighthood, conferred in 1942, the Companion of the Order of the Bath, awarded in 1941, Fellowship of the Royal Society, the U. S. Medal for Merit, the Valdemar Poulsen Gold Medal of the Danish Academy of the Technical Sciences, and the Hughes Medal of the Royal Society for his pioneer researches in radio-telegraphy.

Sir Robert, who is now engaged in work on the peacetime applications of radar, especially in the service of civil aviation, became a Senior Member of IRE in 1946, and a Fellow in 1947. He is President of the Royal Meteorological Society and of the Institute of Navigation.

The value of IRE papers, in a large measure, depends on the ease with which they can be assimilated by the reader. Prospective authors are urged to give thoughtful consideration to suggestions here given to that effect.

The following guest editorial is by a member of the IRE Board of Editors, who is also Vice-President in Charge of Engineering of the United Broadcasting Company, and President of Cleveland Institute of Radio Electronics. As the author of the radio and communication engineering text material used by that school, and as a practicing engineer, his proposal should be helpful to prospective authors of technical papers.—*The Editor.*

Technical Writing for Students

CARL E. SMITH

During the many years I have been preparing technical material for broadcasting employees, and for student consumption at a school, I have come to the conclusion that such material must be carefully organized and skillfully presented if it is to have a real teaching value. The content must speak the truth, of course, but no matter how great the professional value of the material presented, this value is lost to the reader unless effectively presented.

This means, first, that the technical writer should properly appraise the mental capacity of his reader audience. For example, in my IRE editing experience, I have found that most technical reports prepared for company records or from postgraduate theses need to be thoroughly revised to make them suitable for IRE publication. In a way, anyone who reads a technical article or paper is a student and he brings to the reading a certain basic technical knowledge. In addition, he has developed a certain degree of ability to absorb new information of a technical nature. I believe the writer can make this absorption process a little easier for him by organizing the subject matter in a step-by-step manner. This procedure should be as much in line with the reader's normal, logical thinking processes as possible, NEVER putting the cart before the horse. To oversimplify is to write down. To stop short of a full exposition imposes an unfair mental burden on the reader. A crystal-clear word description profusely illustrated with diagrams is to be preferred in qualitative treatments. Even in quantitative writing it is preferable to minimize the mathematics in the text and clearly state in words the significance of each step, relegating intermediate mathematical steps to an appendix when necessary. In many cases a word description with references and the final design equation will suffice. To make sure the reader understands the subject matter thoroughly, it should be illustrated with practical examples.

When I write a technical paper for publication or a lesson for our students or technical broadcasting employees, my procedure in general is to develop the theory inductively, select appropriate illustrations, and if necessary develop design equations step by step from the fundamentals. Then the subject matter is reduced to practice with examples. When I receive a troubled query from some reader or student I feel I have failed to teach with full effectiveness or my method of presentation is at fault, and I conclude that a re-write is justified. We can do that with lessons, but not with technical articles or books. So, in the case of a writer preparing a technical paper or article I suggest that the script be read by the student type of reader before going to the editor. That type of reader is the most appreciative and also the most critical. If you satisfy him you will also be doing a better job for the other type, the casual reader.

Surely an engineering article should teach engineering. I firmly believe that writers on an engineering level should try more earnestly to instruct as well as describe and expound; then more good will be accomplished by our articles.

Report on the International Television Standards Conference*

I. Introduction

The International Radio Consultative Committee (CCIR) held a conference at Zurich, Switzerland, on July 4–14, 1949, to study international problems relating to television. This conference, under the sponsorship of CCIR Study Group 11 (Television), was called in accordance with recommendations of the CCIR Plenary Session held at Stockholm, Sweden, in 1948. The results of the conference will be considered and acted upon at the next Plenary Session to be held in Prague in the spring of 1951. It is hoped that these deliberations will result in international agreements on television standards which will minimize interference between stations, facilitate the interchange of programs among nations, and provide a television system based on the best technical knowledge available.

Prior to the conference, a questionnaire was sent to the administrations of all nations associated with the CCIR. Replies to the questionnaire were received from thirteen nations and three commercial companies. Of these, eleven nations and three commercial companies sent delegates to the conference, as follows: Austria, Belgium, Denmark, U.S.A., France, Italy, Netherlands, Great Britain, Sweden, Switzerland, Czechoslovakia, Cie. Gen. de T.S.F. (France), L. M. Ericsson (Sweden), and RCA (U.S.A.). Hungary and Yugoslavia replied to the questionnaire but were not represented at the conference.

II. National Positions on Television Standards

A. The United States

The American position was that the U.S.A. television standards (525 lines, 30 frames, 4.25-Mc bandwidth) are the most suitable for adoption as a worldwide or regional standard on the basis that they would provide a high quality of service with a minimum of restriction on future progress.

B. Great Britain

Great Britain advocated the adoption of the British standards of 405 lines and 2.75-Mc bandwidth on the basis that they provided a satisfactory quality of service. Moreover, the British government is committed to continue service on these standards domestically for a number of years. It might therefore be necessary for neighboring European nations to adopt the 405-line standard in order to exchange programs with Great Britain. Furthermore, in order to minimize interference, the stations on the continent should adopt carrier frequencies, in the band 41 to 68 Mc, identical to the British frequencies. This allocation involves one 6.75-Mc channel for double sideband operation and four 5-Mc channels for vestigial sideband operation.

C. France

The French government operates at present a 450-line service and, in addition, has committed itself to institute in the future an 819-line service using a 10.4-Mc bandwidth and a 14-Mc channel. During the meeting, France agreed to change from the 450-line service to the British standard of 405 lines. On higher bands (174–216 Mc and 470–960 Mc), they intended to use the 819-line standard for monochrome service with the hope of finding a method of converting from 405-line images to 819-line images in order to facilitate the interchange of programs between their two domestic services, as well as between French and British services.

D. Other Nations

The other nations, headed by Netherlands and Sweden, favored a 625-line 25-frame standard on a bandwidth of 4.75 Mc and a channel of 6.75 Mc. They felt that, on the one hand, the British 405-line standard was "obsolescent" or "low-definition," and that, on the other hand, the U.S.A. 30-frame standard was unsuitable for 50-cps power systems. The American delegation pointed out that the proposed 625-line 25-frame standard was not in fact very different from the 525-line 30-frame standard so far as equipment operation is concerned, provided that the system is divorced from dependence on the power supply frequency.

III. Conclusions of the Conference

During the conference, unanimous agreement was reached on the following four points:

1. The aspect ratio should be 4 units horizontally and 3 units vertically. This agrees with the U.S.A. standard, but represents a change in the British standard of 5 units by 4 units. The British delegates stated that the BBC has requested permission from the proper government committee to change to the 4 by 3 standard in the near future.

2. Interlacing should be used at a ratio of 2-to-1. This was not a controversial issue.

3. The vertical scanning frequency (number of fields per second) should not be synchronized with the power frequency. Agreement was reached on this important point due to the realization that, first, power systems would not usually be interconnected between nations or even between cities within some nations. Thus nonsynchronous operation would facilitate the interchange of programs between nations and cities not so interconnected. Secondly, nonsynchronous operation would permit the 525-line standard and the 625-line standard to exist compatibly on a worldwide basis.

4. The direction of polarization need not be standardized, as it is not a basic issue. (Six nations favored horizontal and two nations vertical polarization.)

Disagreement persisted on the remaining points. On lines and frames, the U.S.A. recommended the 525–30 value, Great Britain and France 405–25 for the 41- to 68-Mc band, France 819–25 for higher bands, and all other nations 625–25. On positive versus negative modulation, 4 nations voted for negative against two (Great Britain and France) for positive, the remaining nations indicating no marked preference. On AM versus FM sound, 7 nations favored FM and two nations (Great Britain and France) AM.

The arguments for negative modulation and FM sound were given additional weight in the light of experience with the intercarrier sound system. The British stated that the intercarrier system could be used with positive modulation (and FM sound) provided the synchronizing pulses were limited at some level above zero carrier. The British synchronizing pulses as now radiated do not reach the zero level. The British announced their intention of experimenting with the intercarrier system, using positive modulation, and promised to report the results to the next meeting of the Study Group. The U. S. delegation invited the Study Group to hold its next meeting in the United States in order that American experimentation in color television and in minimizing interference between stations, as well as other matters on the agenda, might be viewed first hand.

IV. Questions Posed for Further Study

In conclusion, the Study Group formulated the following questions as needing further study:

Question 1.

What are the technical requirements and the additional cost necessary to permit the operation of a television system to be independent of the frequency of the power supply?

Question 2.

What is the relation between maximum brightness and field frequency for absence or flicker, for field frequencies of the order of 50 to 60 fields per second, taking into account the decay time of the different colors?

Question 3.

What is the value, in minutes of angle, of the resolving power of the eye in photographic and in television systems for ranges of brightness values and distances of viewing most commonly met with in television?

Question 4.

What are the conditions of measurement required to allow the comparison of the respective qualities of television and cinema pictures, and what are the results of such comparisons?

Question 5.

What are the respective advantages and disadvantages of positive and negative picture modulation and of amplitude and frequency sound modulation, taking into account the following four points:

(a) Effect of noise on picture
(b) Effect of noise on synchronization
(c) Automatic gain control
(d) Intercarrier sound reception?

Question 6.

What are the possibilities of standardization and what are the preferred characteristics in the recording of television pictures on film?

* Decimal classification: R007.9×R583. Adapted from "Report to the RMA Engineering Department on the International Television Standards Conference held at Zurich, Switzerland, July, 1949," by Donald G. Fink.

Conductive Plastic Materials*

MYRON A. COLER†, F. ROBERT BARNET‡, ALBERT LIGHTBODY‡, AND H. A. PERRY, JR.‡

A NEW CLASS of plastic materials has been developed. The outstanding characteristic of these materials resides in the fact that they have substantial and predeterminable electrical conductivities, and yet possess the desirable general mechanical and fabrication properties of ordinary plastics. Thermosetting, thermoplastic and elastomeric variants have been produced.

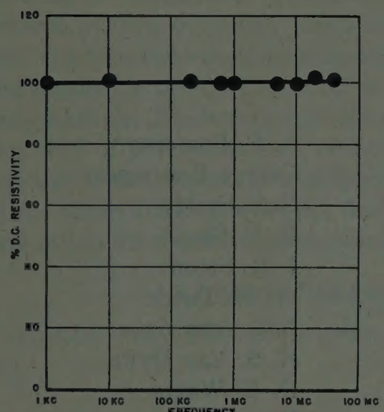

Fig. 1—Effect of frequency on resistivity. Markite MS-501 at 24°C.

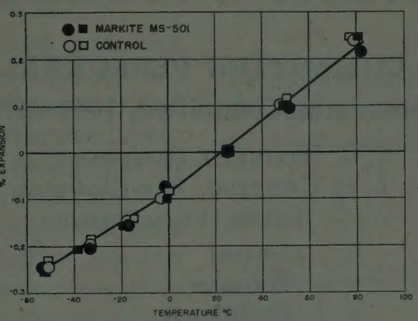

Fig. 2—Thermal expansion versus temperature. Reference temperature, 25°C.

Representative data for one type are presented in Table I. Particular attention is called to the significant thermal and electrical conductivities; the latter is approximately 10^{12} times that of a general purpose phenolic. The material follows Ohm's Law at low and moderate current densities and appears to be free of frequency variation effects (see Fig. 1). The temperature coefficient of resistivity is positive and of the order of 0.2 per cent/°C.

It has proven possible to match the thermal expansion of the MS-501 material with an insulator plastic throughout the range from −50°C to 80°C (see Fig. 2). Since the materials are comoldable, this favors the molding of composite conductor-insulator structures which will be liquid-tight and nonseparating at extremes of temperature.

The range of resistivities covered by these conductive plastic materials is indicated in Fig. 3. The range extends roughly from the resistivity of distilled water to values approaching that of mercury. Many of the materials can be metal-plated directly. As a group, they exhibit comparatively low densities which commends their use where weight reduction is important, e.g., on aircraft and in devices to be rotated at great speeds. Other more specific applications of interest include low ohmic resistors for the higher frequencies, nonlinear variable potentiometers, shields, attenuators, electrodes, and three-dimensional circuits.

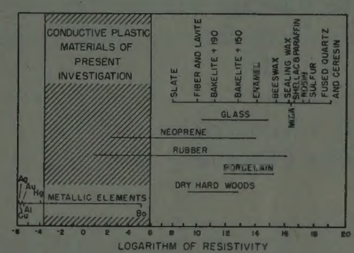

Fig. 3—Resistivity spectrum. Shaded area corresponds to markite materials superimposed. (Based on modification of chart by Yerzley, *Ind. Eng. Chem.*, vol. 35, p. 330; 1943.)

* Decimal classification: R281. Original manuscript received by the Institute, February 28, 1949; abstract received, August 26, 1949.

† Markite Company, New York, N.Y.
‡ Plastics Division, Naval Ordnance Laboratory, White Oak, Md.

TABLE I

MARKITE MS-501 CONDUCTIVE PLASTIC MATERIAL
Typical Data on Physical Properties

1	Type	Thermosetting
2	Fabrication Methods	Compression, Molding, Machining, Grinding
3	Color	Black
4	Specific Gravity	1.57
5	Tensile Strength—Ultimate, *psi*	7,400
6	Elongation, %	0.80
7	Tensile Modulus of Elasticity, $psi \times 10^{-5}$	13
8	Tensile Strength—Proportional, *psi*	2,100
9	Flexural Strength—Ultimate, *psi*	11,000
10	Flexural Modulus of Elasticity, $psi \times 10^{-5}$	11
11	Flexural Strength—Proportional, *psi*	7,700
12	Compressive Strength, *psi*	26,400
13	Impact Strength—Izod ft. lbs./in. notch	0.37
14	Hardness—Rockwell, *M*	106
15	Moisture Vapor Trans., g/100 sq. in./24 hr./mil	44
16	Water Absorption gain, %	0.55
17	Specific Heat, Cal/°C/g	0.31
18	Thermal Expansion Coefficient, cm/cm/°C$\times 10^5$	3.8
19	Thermal Conductivity, cal/sec/cm²/C$\times 10^4$	21
20	Heat Distortion Temperature, °F	306
21	Electrical Resistivity, ohm-cm	0.01

Standards on
DESIGNATIONS FOR ELECTRICAL, ELECTRONIC AND MECHANICAL PARTS AND THEIR SYMBOLS, 1949*

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1. INTRODUCTION

It is the purpose of this Standard to establish general principles governing the formation and application of reference designations and to provide a list of such designations applicable to parts used in electrical and electronic applications.

1.1 Reference Designations

Reference designations are combinations of letters and numbers used to identify the parts of a piece of apparatus or equipment on schematic (see par. 2.4) and other drawings, diagrams, parts lists, instruction books, etc. The reference designations may also be stamped on the apparatus or equipment on or near the parts which they identify. The letters in a reference designation show the kind of part, such as a resistor, amplifier, electron tube, etc., and the number differentiates between parts of the same kind. A reference designation is not an abbreviation for the name of the part.

1.2 Other Designations

This Standard does not prevent the use of other than reference designations. In cases where other designation are used, it is preferable that the reference designations be used in addition. The other designations will be underlined except when confusion between the two types of designations will not result.

1.3 Scope

This Standard supplements and is parallel with *Standards on Abbreviations, Graphical Symbols, Letter Symbols and Mathematical Signs, 1948.*

1.4 Purpose

Reference designations have two main purposes as follows:

1.4.1 For reference to parts in descriptions of the operation of a circuit or mechanical device.

1.4.2 For the identification of replaceable parts in stock control and maintenance operations.

2. DEFINITION OF TERMS

2.1 Code

For this Standard a code generally consists of a name (which will usually indicate a function) plus one of the following: A number, drawing number, model number, specification number, or style number, etc. These numbers consist of numerals and letters alone or in various combinations. The name portion of a code may be omitted when the meaning as conveyed by the graphical symbol or reference designation is clear. For example, "6AK5" is usually used in conjunction with the proper graphical symbol instead of "6AK5 TUBE."

Examples:

Number+Name	—913 Modulator
Drawing Number+Name	—ED-64372-01 Waveguide Assembly
Name+Model Number	—Tuning Unit, Model 519D
Specification Number+Name	—J6864 Amplifier
Name+Style Letters	—Generator, SK

2.2 Major Assembly

For this Standard, a major assembly is a complete piece of apparatus or equipment as supplied by a manufacturer and identified by a code. The term "assembly" is often used instead of "major assembly." For the purpose of applying reference designations, a major assembly will usually be the largest unit employing a single system of reference designations. A major assembly is composed of parts which may be either elements or subassemblies, coded, or uncoded.

2.3 Subassembly

2.3.1 For this Standard a subassembly is a part of a major assembly consisting of a group of parts which, for purposes of identification, is considered as a unit.

2.3.2 A subassembly may or may not be replaceable as a unit. A grouping of parts indicated by boundary lines on a schematic diagram is not necessarily an integral physical unit. Since the schematic diagram is primarily functional, the grouping of parts is also functional and may include parts which in the actual equipment are separated physically.

2.3.3 A subassembly will usually be coded, but coding is not a necessary criterion for a subassembly.

2.3.4 A subassembly may be given a reference designation and the parts of a subassembly may also be given reference designations.

2.3.5 Examples of subassemblies are: Amplifier, filter, printed circuit, rectifier, and waveguide assembly.

2.3.6 Subassemblies which require separate identification on schematic drawings in accordance with the principles outlined herein will be circumscribed by a line which may be either:

- - - - - - - - - - -, indicating an electrical shield which is also a mechanical boundary,

or

——— — — ———, indicating the boundary of a mechanical grouping or a grouping for convenience.

The presence of a shield line around a group of apparatus does not necessarily indicate that it is a subassembly for the purposes of this Standard.

2.4 Schematic Drawings

In this Standard "schematic drawing" includes drawings known as "Elementary Diagrams" in the power field.

3. FORMATION OF REFERENCE DESIGNATIONS

3.1 Letters

3.1.1 The letter or letters of the reference designation shall be as indicated in par. 7.1.

3.1.2 The letter *E*, in the case of terminals, may be omitted from drawings and not stamped on equipment for such parts as transformers, panels, sockets, etc. where no confusion will result.

3.2 Numbers

The number of the reference designation follows the letter or letters without a hyphen and shall be of the same size and on the same line. For example C1, S14 and MG5. The assignment of numbers should preferably start with 1 in the upper left-hand corner of the schematic drawing of the major assembly or subassembly and proceed in numerical order in a logical manner through the drawing. However, it is not required that the series of numbers be necessarily consecutive or complete. For instance, in the case of successive improvements of a piece of equipment, some of the parts may be eliminated. Here it is unnecessary to redesignate the remaining parts merely to keep the number series consecutive. The parts list will show which numbers are missing.

4. APPLICATION OF REFERENCE DESIGNATIONS

4.1 General

4.1.1 Reference designations are assigned to electrical and mechanical parts on schematic and other diagrams and in instruction books, spare parts lists, etc. Common hardware items such as bolts, nuts, and washers will usually not require reference designations.

4.1.2 The assignment of reference designations will usually be made on the schematic drawing, which consequently controls the designations to be applied on other drawings, such as equipment and assembly drawings, and in parts lists.

4.1.3 Reference designations should also be marked on or near the appropriate parts. It is recognized that in some cases, particularly in compact groupings of small parts, this will be impracticable.

4.1.4 When a reference designation, other designation or a code appears on a schematic diagram near a graphical symbol and it is *NOT* intended that it be marked on or near the actual part, such designation or code shall be enclosed in parentheses on the schematic diagram.

4.1.5 The assignment of the number portions of the reference designations for the parts will depend on the conditions of manufacture, supply and use. Two cases can be distinguished: (1) A major assembly with a single number series; and (2) a subassembly with its own internal number series.

4.2 Major Assembly with a Single Number Series

In this case a single series of numbers shall be used for each kind of part throughout a major assembly, regardless of whether the parts are included in subassemblies or not. This method will in general be confined to the smaller and simpler major assemblies, usually those without separately manufactured subassemblies. Fig. 1 is an example. For purposes of illustration only, this major assembly has been coded an XX Major Assembly. This same major assembly, with changes in the assumptions as to how it is subdivided, is used for the other examples of the methods of designating parts of subassemblies. Fig. 2 is an example in which a major

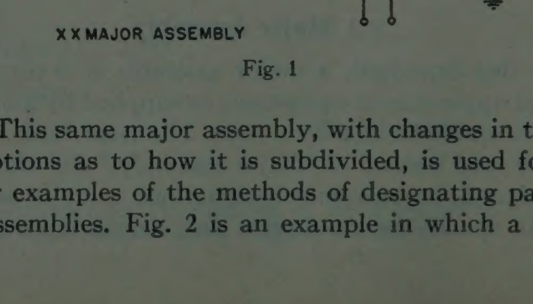

Fig. 1

assembly has been divided into two subassemblies and a remainder. The parts designations are here identical with those of Fig. 1. In this case the reference designations for the subassemblies themselves may be convenient for description, and for correlation with block diagrams, such as Fig. 3, but they are not essential for the identification of a part within a subassembly since the part designation is itself sufficient.

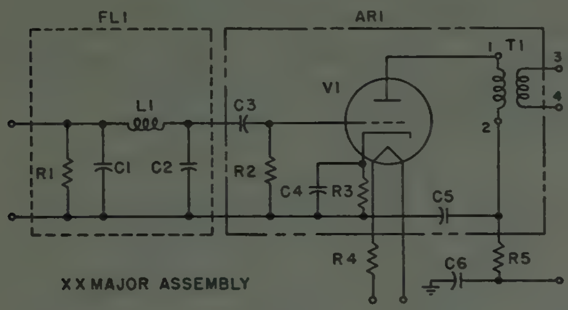

Fig. 2

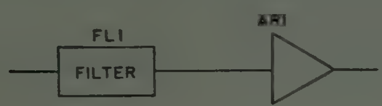

Fig. 3

4.3 Subassembly with Its Own Internal Number Series

There is a case in which it is advantageous to number the parts of a subassembly with its own internal number series. This case is especially applicable to multiple-use subassemblies which may appear more than once in a major assembly, or in more than one major assembly. In this case the reference designations internal to the subassembly will remain the same, no matter where the subassembly is used, and the code of the subassembly shall appear on the pertinent drawings. In this case the parts of a major assembly not included in the above subassemblies (but including parts in subassemblies not assigned separate internal number series) shall be numbered in a single series as outlined in par. 4.2. Fig. 4 illustrates this case.

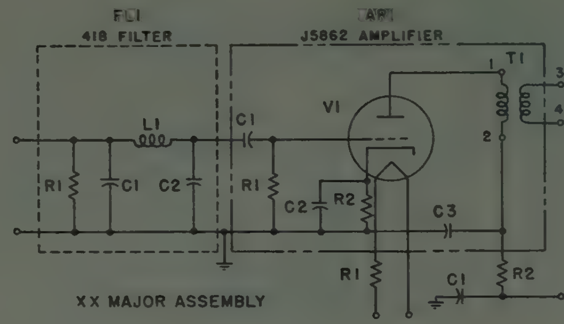

Fig. 4

The parts of subassemblies having separate number series may be identified either by means of the code of the subassembly and the part reference designation or by means of a compound reference designation formed by prefixing the part designation with the subassembly designation.

5. COMPLETE IDENTIFICATION OF PARTS

5.1. Abbreviation of Compound Reference Designations

Under certain conditions where no confusion will result, the initial letter part may be omitted from a compound reference designation. This provision applies to such cases as are exemplified by Fig. 5. It should be noted that in order to apply this principle the number portions of the reference designations for any two subassemblies can not be the same even though the subassemblies are of a different kind.

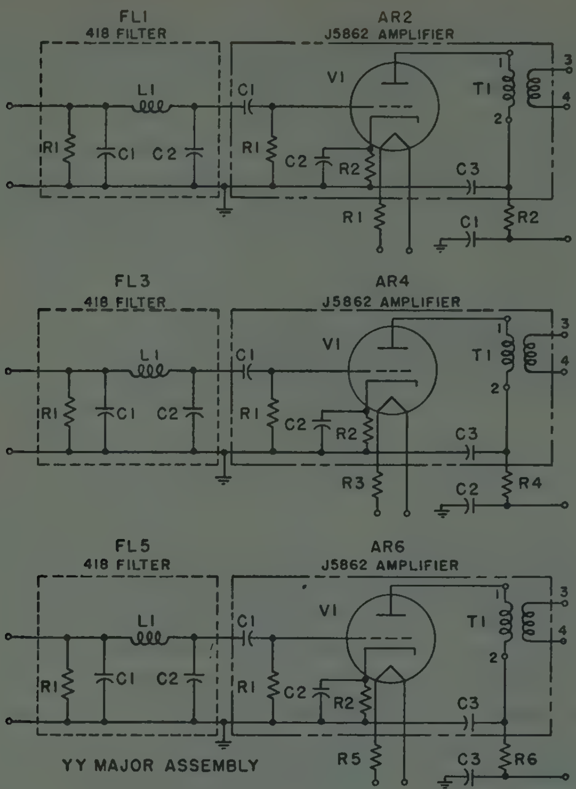

Fig. 5

5.2 Examples of Compound Reference Designations

Examples of compound reference designations and abbreviated compound reference designations are given in the table below. Where applicable (the case of Fig. 4) examples are given of parallel identification of the same parts by the use of the subassembly code, together with the reference designation for the part.

TABLE I

EXAMPLES OF IDENTIFICATION OF PARTS (NOTE 1)

Identification of same part in			
Fig. 1	Fig. 2	Fig. 4	Fig. 5 (note 2)
R1	R1	(a) FL1R1 (b) R1 of 418 Filt.	1R1, 3R1, 5R1
L1	L1	(a) FL1L1 (b) L1 of 418 Filt.	1L1, 3L1, 5L1
C4	C4	(a) AR1C2 (b) C2 of J5862 Ampl.	2C2, 4C2, 6C2
R3	R3	(a) AR1R2 (b) R2 of J5862 Ampl.	2R2, 4R2, 6R2
R4	R4	R1	R1, R3, R5

Note 1. The code XX Major Assembly (or YY Major Assembly for Fig. 5) is a necessary part of the complete identification except for the examples (b) of Fig. 4 where the code, 418 Filter or J5862 Amplifier, is used in its place.

Note 2. Since the parts of Fig. 1 appear three times in Fig. 5, 3 entries of each are given. All are examples of abbreviated compound reference designations.

6. SPECIAL PROCEDURES FOR SOCKETS AND CONNECTORS

6.1 Sockets

A socket which is always associated with a particular plug-in or screw-in device, such as an electron tube or a fuse, should be designated by a symbol which includes the symbol of the inserted device. For example, the socket for fuse F7 should be designated XF7.

6.2 Connectors

In the case of a stationary connector, the reference designation both of the connector and of the associated movable connector or adapter (this latter in parentheses) should be marked on the chassis or panel adjacent to the stationary connector.

Examples: a. J5 b. J5
(P5) (CP4)
(P4)

In this case the reference designation in parentheses will not usually be found on the schematic drawings.

7. LISTS

7.1 Alphabetically by Part Name

7.1.1 Classes of parts included in par. 7.1 have been marked with an asterisk to facilitate the designation of parts not specifically included in the list. In case of doubt, a letter or letters already assigned to the part or class most similar in function should be used.

Item

AR	Amplifier
E	Antenna
E	Arrester, lightning
AT	Attenuator
B*	Blower, fan, motor, prime mover
T	Autotransformer
BT	Battery
I	Bell
TB	Block, connecting
CB	Breaker, circuit
I	Buzzer
W	Cable
C	Capacitor
V	Cell, light-sensitive, photo-emissive

15	L	Choke
	N	Chart
	P	Connector (movable portion or portion located on a plug-in device)
	J	Connector (stationary portion)
	E	Contact, electrical
20	K	Contactor (electrically operated)
	S	Contactor (mechanically or thermally operated)
	M	Counter (indicating device)
	DC	Coupler, directional
	CP	Coupling (aperture, loop or probe)
25	E	Counterpoise
	HY	Coil, hybrid
	L	Coil, induction, loading, relay operating, retardation, tuning
	T	Coil, repeating
	Y	Crystal, piezoelectric
30	S	Cut-out, thermal
	CR	Detector, crystal
	I*	Device, indicating (except meter or thermometer)
	S	Dial (circuit interrupter)
	I*	Dial or indicating device (except meter or thermometer)
35	DP	Diaphragm
	D	Dynamotor
	EQ	Equalizer
	G	Exciter
	B	Fan
40	FL	Filter
	F	Fuse
	M*	Gauge, meter, thermometer, etc.
	G	Generator
	HS	Handset
45	H*	Hardware, bolts, nuts, screws, etc.
	PU	Head, erasing, recording, reproducing
	HT	Headset, or telephone receiver (not part of handset)
	HR	Heater (element for thermostat, oven, etc.)
	L	Inductor
50	P	Jack (connector, movable portion or portion located on a plug-in device)
	J	Jack (connector, stationary portion)
	HY	Junction, hybrid
	CP	Junction, Tee or Wye
	S	Key-switch
55	S	Key, telegraph
	RT	Lamp, ballast
	HR	Lamp, heating
	I	Lamp, illuminating
	RT	Lamp, resistance
60	I	Lamp, signal
	Z	Line, artificial
	DL	Line, delay
	LS	Loudspeaker
	M	Meter
65	MK	Microphone
	Z	Modulator
	B	Motor
	MG	Motor-generator
	A*	Mounting not in electrical circuit (not a socket)
70	N*	Nameplate, chart, etc.
	Z	Network, general (where specific letters do not fit)
	Y	Oscillator (excluding electron tube used in oscillator), magneto-striction
	AT	Pad
	E*	Part (misc. electrical)
75	U*	Part, hydraulic
	O*	Part (misc. mechanical), bearing, coupling gear, shaft, etc.
	A*	Part, structural
	W*	Path, guided transmission
	PU	Pickup, erasing head, recording head, reproducing head
80	P	Plug (connector, movable portion or portion located on a plug-in device)
	J	Plug (connector, stationary portion)
	R	Potentiometer
	PS	Power-supply
	B*	Prime-mover
85	E	Protector (carbon-block or gap)
	RE	Receiver, radio
	HT	Receiver, telephone (not part of handset)
	J	Receptacle (connector, stationary portion)
	CR	Rectifier, crystal or metallic
90	VR	Regulator, voltage (except an electron tube)
	M	Recorder, elasped time (clock)
	M	Register, message
	K	Relay (electrically operated switch)
	RP	Repeater (telephone usage)
95	R	Resistor
	R	Rheostat
	I	Ringer, subscriber's set
	E	Shield, electrical
	E	Short
100	X	Socket (see paragraph 6.1. Examples: XAR XF, XV, etc.)
	L	Solenoid
	LS	Speaker, loudspeaker
	TB	Strip, terminal
	A*	Structural part
105	Z	Subassembly, general (where specific letters do not fit)
	S	Switch (mechanically or thermally operated)
	E	Terminal, individual
	AT	Termination, resistive
	M	Thermometer
110	RT	Thermistor
	TC	Thermocouple
	S	Thermostat
	MT	Transducer, mode
	T	Transformer

115 Q Transistor
MK Transmitter, telephone
TR Transmitter, radio
V Tube, electron
CR Varistor
120 I Vibrator, indicating
W Waveguide
L Winding, relay
W Wire

7.2 List—Alphabetically by Letters

7.2.1 Classes of parts included in par. 7.2 have been marked with an asterisk to facilitate the designation of parts not specifically included in the list. In case of doubt, a letter or letters already assigned to the part or class most similar in function should be used.

Item

A* Mounting (not in electrical circuit and not a socket), structural part
AR Amplifier
AT Attenuator, pad, resistive termination
B* Blower, fan, motor, prime mover
5 BT Battery
C Capacitor
CB Circuit breaker
CP Coupling (aperture, loop, or probe), coaxial or waveguide junction (Tee or Wye)
CR Crystal detector, crystal or metallic rectifier, contact rectifier, varistor
10 D Dynamotor
DC Directional coupler
DL Delay line
DP Diaphragm
E* Antenna, binding post, counterpoise, dipole antenna, electrical shield, electrical contact, electrical parts (misc.), lightning arrestor, individual terminal, loop antenna, protector (carbon-block or gap), short
15 EQ Equalizer
F Fuse
FL Filter
G Exciter, generator, vibrator (rectifying)
H* Hardware, bolts, nuts, screws, etc.
20 HR Heater, (element for thermostat, oven, etc.), heating lamp
HS Handset
HT Telephone receiver (not part of handset), headset
HY Hybrid coil or hybrid junction
I* Bell, buzzer, dial, illuminating lamp, indicating device (except meter or thermometer), signal lamp, subscriber's set ringer, vibrator (indicating)
25 J Connector (stationary portion), jack, plug, receptacle
K Relay (electrically operated contactor or switch)
L Choke, inductor, loading coil, induction coil (telephone usage), relay operating coil, retardation coil, solenoid, tuning coil
LS Loudspeaker, speaker
M* Counter (indicating device), elapsed-time recorder (clock), gauge, message register, meter, thermometer
30 MG Motor-generator
MK Microphone, telephone transmitter
MT Mode transducer
N* Chart, nameplate, etc.
O* Mechanical part, bearing, coupling, gear, shaft, etc.
35 P Connector (movable portion or portion located on a plug-in device), jack, plug
PS Power supply
PU Pickup, erasing head, recording head, reproducing head
Q Transistor
R Potentiometer, resistor, rheostat
40 RE Radio receiver
RP Repeater (telephone usage)
RT Ballast lamp, resistance lamp, thermistor
S Contactor, dial (circuit interrupter), key-switch, telegraph key, mechanically or thermally operated switch, thermal cut-out, thermostat
T Autotransformer, induction coil (telephone usage), repeating coil (telephone usage), transformer
45 TB Connecting block, group of individual terminals on its own mounting, terminal board, terminal strip
TC Thermocouple
TR Radio transmitter
U* Hydraulic part
V Electron tube, light-sensitive cell, photo-emissive cell
50 VR Voltage regulator (except an electron tube)
W* Cable, coaxial cable, guided transmission path, waveguide, wire
X Socket (See par. 6.1. Examples XAR, XF, XV)
Y Oscillator (excluding electron tube used in an oscillator), piezoelectric crystal, magneto-striction oscillator
Z* Network or subassembly, general (where more specific letters do not fit), artificial line, modulator

A Microwave System for Television Relaying*

J. Z. MILLAR†, SENIOR MEMBER, IRE, AND W. B. SULLINGER†

Summary—**The transmission requirements for radio relay systems for television network operation are discussed. A system designed by the Philco Corporation and employing heterodyne modulation with a SAC-19 Klystron, developed by the Sperry Gyroscope Corporation for this application, is described. Western Union has installed this equipment, which operates in the 6,000-Mc common carrier band, between New York and Philadelphia, and photographs are included showing the repeater, antenna, and the results of a square-wave and CBS test pattern after transmission over the circuit.**

PIONEER work in the application of microwaves to telegraphy was done by the Western Union Telegraph Company developing jointly with the Radio Corporation of America an experimental relay system between New York and Philadelphia in 1945. This circuit was later expanded by Western Union to include a second circuit via an alternate route between these two cities. Also, a triangular system connecting New York, Pittsburgh, and Washington has been completed.[1] It was envisioned from the beginning that the same physical facilities, towers, buildings, power plants, and the same maintenance personnel which were used for the telegraph circuits, could be shared with television systems. Accordingly, in 1947, active development work was initiated with the Philco Corporation for the design of equipment capable of meeting the exacting requirements of television network operation.

In an attempt to define the transmission objective, it is assumed that a practical value of signal-to-noise ratio in a video channel which has been relayed from New York to San Francisco would be 25 db (measured at the peaks, or maximum instantaneous amplitudes, of signal and noise). This value is far from the ideal value—it is the minimum value considered useful. Since interconnection of similar systems is planned to facilitate networking, 9 db should be added, and also a 6-db maintenance margin. This increases the system requirement to 40 db between major cities which are considered to be logical switching points, a value readily realizable in current microwave practice.

Of course, it would not be possible to devote enough frequency spectrum to each video program channel to permit the use of a subcarrier and double frequency modulation[2] as was done in the RCA system[3] for telegraph transmission. However, one level of frequency modulation is feasible if the deviation index is held to a value near unity, but the FM noise improvement will then be only 5 to 10 db. This transmission method would require a radio-frequency band of approximately 20 Mc to accommodate the sidebands, and a minimum peak carrier-to-noise ratio of 35 db will be required for an intercity system having approximately 10 repeaters. This will require a signal level on the input to each intermediate-frequency system of approximately 750 microvolts.

As it is impractical to use diversity reception in a system which does not utilize a subcarrier, and especially since video signals are sensitive to phase distortion, the amount of margin needed to overcome the effects of fading will be at least 30 db, and might be as high as 40 db, depending upon the continuity of service desired. Assuming space losses corresponding to normal repeater station separations, and conventional antenna power gains, the required transmitter power will be approximately 15 watts, if it is assumed that fading does not occur in all repeater sections simultaneously. The choice of a microwave frequency band for television relay is dictated by many factors. The 4,000-Mc band is more stable than the 6,000-Mc band with respect to fading and atmospheric attenuation, but requires much larger and more costly antennas for the same power gain and larger waveguides and support structures. An examination of propagation records, considering the hours most useful for broadcasting, leads to the conclusion that the 6,000-Mc band would be satisfactory for television transmission. The common carrier band between 10,700 and 11,700 Mc is considered primarily useful only for short-distance relaying.

The most controversial parameter of a video transmission system is the frequency response required for broadcasting networks. Perhaps it is a little forward-looking to provide a video response ranging from 30 cycles to 5 Mc, when none of the receivers being used can be manufactured to respond to frequencies above about 4 Mc because of the required separation between the video intermediate-frequency and sound intermediate-frequency frequencies of 4.5 Mc. Nevertheless, this response should be provided to take care of future requirements, and to permit transmission of undistorted synchronizing pulses.

It is a fortunate circumstance that amplitude distortion is not very noticeable in a television image. A circuit having a combined harmonic production when tested with a single frequency of less than 5 per cent (down 26 db) would probably pass unnoticed by the average observer. In contrast, the relative phase relations of the frequency components of the complex television wave must be preserved by passing the signals only through circuits having almost linear phase shift; that is, circuits where the phase shift is proportional to frequency. A departure from phase linearity, that is, a delay distortion of as much as 50 milli-microseconds,

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[1] J. Z. Millar, "A preview of the Western Union system of radio beam telegraphy," *Jour. Frank. Inst.*, vol. 241, no. 6, June, 1946; also vol. 242, no. 1, July, 1946.

[2] Leland E. Thompson, "A microwave relay system," PROC. I.R.E., vol. 34, pp. 936–942; December, 1946.

[3] G. G. Gerlach, "A microwave relay communication system," *RCA Rev.*, vol. 7, pp. 567–601; December, 1946.

is objectionable in a television image. Since this is an over-all requirement, it can be seen that the phase distortion requirement for each intercity system, and to a greater degree for each repeater of a relay system, will be exceedingly rigorous. Furthermore, the design of all component parts of the equipment must be co-ordinated, and phase equalization applied wherever practicable.

Such perfection in phase response requires the employment of testing and alignment techniques making use of transient, or time-function, methods. These techniques are also invaluable in the laboratory. It is further desirable constantly to monitor terminal-to-terminal transmission with high-quality monitors to detect gradual signal quality deterioration before it can become serious.

The most obvious method of accomplishing the repeatering function is to demodulate, amplify, and then modulate a new carrier frequency. The difficulty with this method is that repeated demodulation and modulation will introduce cumulative distortion because of nonlinearity in the functional units.

Of course, by using antennas of good enough front-to-back ratio, a straight-through radio-frequency repeater is possible, but present experience indicates that all tubes capable of amplifying microwave frequencies are too noisy for this application, except at very short repeater spacings. The best choice of radio repeater, if a subcarrier type transmission system cannot be employed, is the so-called heterodyne repeater. In this repeater, the incoming wave is translated by heterodyning to the intermediate-frequency range, amplified, and then translated back to the microwave range, and further amplified before being radiated. This system most closely approaches the transmission goal, but because of equipment limitations it may be some time before coast-to-coast circuits meeting the minimum bandwidth and distortion requirements will be available.

The term "heterodyne modulation" describes any system by which radio-frequency carrier is mixed with a modulated wave to produce modulated rf sidebands. One system for accomplishing this result makes use of a balanced crystal mixer, but because of power limitations of the crystals, several stages of microwave klystron amplification were required. The Sperry Type SAC-19 tube, known as a heterodyne-mixer and developed by the Sperry Gyroscope Corporation for this application, accomplishes the desired result in one stage, and produces a power output of from 1 to 1.7 watts. Fig. 1 shows the Sperry SAC-19 tube. It is considered a unique contribution to the art.

For applications where adequate signal-to-noise ratio cannot be obtained with approximately 1.7 watts output, later equipment and an improved version of the tube capable of producing an output power of over 6 watts may be used.

Other design requirements relate to (1) reversibility of circuits to meet more nearly the transmission needs of broadcasters; (2) common antennas and waveguides for all circuits at a repeater or terminal station; and (3) location of all radio components in the building at the base of towers, or in rooms near the top of concrete towers.

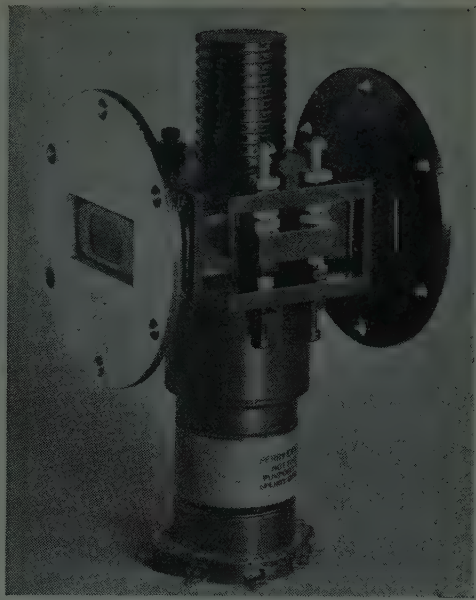

Fig. 1—Sperry SAC-19 tube.

Keeping the above objectives in mind, the two television relay circuits which have been installed between New York and Philadelphia by Western Union will be described. Philco Model TLR-2 Television Radio Relay Equipment was employed.

The two circuits are independently reversible with a switching time of less than 15 seconds, which makes possible the use of a single relay circuit for transmitting consecutive programs in opposite directions, and reduces the amount of equipment necessary as compared to two-circuit service.

The bandwidth required for each transmitter is 20 Mc and adjacent channels are separated by 40 Mc between center frequencies. Alternate stations in a given direction of transmission use alternate frequency bands. Therefore, the two reversible circuits between New York and Philadelphia require only four properly spaced 20-Mc bands.

The operation of a typical repeater can best be described by reference to the block diagram shown in Fig. 2. In a typical repeater installation, the incoming 6,115-Mc signal is picked up on the antenna and is fed down the 100-foot, $1\frac{1}{2}$- by $\frac{3}{4}$-inch waveguide, to the antenna switch. Following the antenna switch is an rf filter for preselection. This filter provides approximately 25 db of selectivity at frequencies 40 Mc removed from the channel. The signal from the filter is coupled through a flexible waveguide into the crystal mixer, which is mounted on the receiver chassis.

The 75-Mc intermediate frequency is obtained by beating the incoming signal against the transmitter local oscillator, which in this case would be tuned to 6,040 Mc. The 75-Mc intermediate-frequency signal obtained from this amplifier is fed through a 50-ohm cable into the converter unit in the transmitter at the relay station. A discriminator is also included at the

end of the 75-Mc intermediate-frequency amplifier, and the output of this discriminator is amplified by a

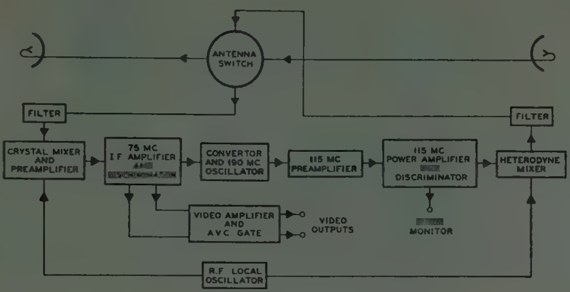

Fig. 2—Block diagram of a repeater.

three-stage video amplifier to provide a 2-volt peak-to-peak video signal for driving both a television transmitter and a video monitor separately. This video amplifier is a necessary part of the terminal receiver, and in a repeater unit it is included both for monitoring purposes and to permit the dropping of a program at intermediate relay points without degradation to through transmission.

The converter unit includes a crystal-controlled oscillator, which is multipled up to 190 Mc. Beating the 75-Mc intermediate-frequency signal against this 190-Mc carrier produces the second intermediate frequency, 115 Mc. The signal is amplified in the 155-Mc preamplifier and in the 115-Mc power amplifier to approximately 80 volts peak-to-peak. A second FM discriminator is provided for monitoring purposes and to provide an AFC signal to the deviator when the same equipment is used as a terminal transmitter.

The 80-volt peak intermediate-frequency signal is connected into the cathode circuit of the SAC-19 tube. This is a two-cavity klystron, tunable throughout most of the 6,000-Mc common carrier band, and which has 20-Mc bandwidth at each cavity setting. When operated as a heterodyne mixer, the buncher cavity of the klystron is tuned to the local oscillator frequency, 6,040 Mc. The catcher cavity of the klystron is tuned to 6,155 Mc, which is the sum frequency of the local oscillator and 115 Mc. The accelerating potential of the klystron is 500 volts, and the 80-volt intermediate-frequency signal modulates the velocity of the electron beam approximately 10 per cent. This varies the transit time of the electrons through the drift space of the klystron, and consequently modulates the energy which is fed into the buncher cavity.

The modulation produces sidebands spaced at 115-Mc intervals from the rf carrier. These sidebands each contain the frequency-modulation intelligence which is present on the 115-Mc intermediate-frequency signal. By tuning the catcher cavity to either the first upper or first lower sideband, a frequency-modulated wave is obtained which contains the desired intelligence. The carrier and the undesired sidebands are suppressed by the selectivity of the catcher cavity, and also by the rf filter which follows the synchrodyne mixer. This filter has a selectivity of approximately 55 db at 115 Mc from the center frequency. A flexible waveguide connects the output of the rf filter into the antenna switch, which in turn connects the transmitter to either the eastbound or the westbound antenna, through the 100-foot waveguides which run up the tower.

Fig. 3 shows the response of the 75-Mc intermediate-frequency amplifier and the discriminator characteristic. A television signal wave is superimposed to indicate the manner in which the frequency is deviated. The terminal transmitter, shown in Fig. 4, includes the same components as the transmitter section of the repeater, with the addition of the deviator, and a deviator power supply. The composite television signal input to the deviator unit is used to deviate two 2K28 reflex klystrons in opposite frequency sense. The outputs of these oscillators are mixed in a crystal to produce a difference frequency of 115 Mc which becomes frequency-modulated in the process in accordance with the input video signal. Maximum deviation, corresponding to the voltage difference between the tip of the synchronizing signal and peak white, causes a deviation of 12 Mc. The difference frequency corresponding to the synchronizing pulse peak is established and held at 123 Mc. One of the deviator oscillators is controlled by an automatic-frequency control circuit in such a way that this frequency reference is held constant, regardless of oscillator drift or picture content. The resulting signal band is then

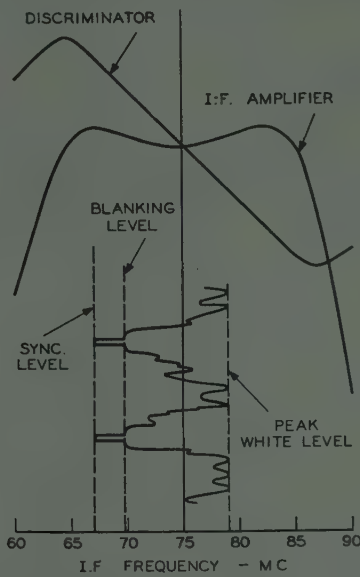

Fig. 3—Characteristics of an intermediate-frequency amplifier and discriminator.

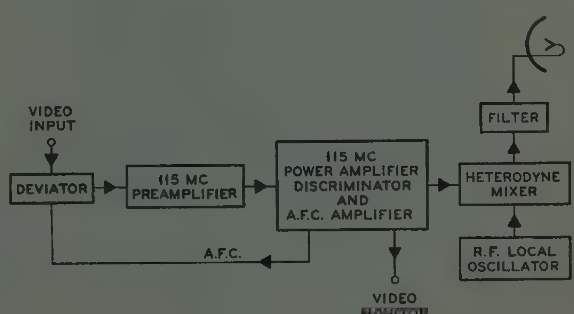

Fig. 4—Block diagram of a terminal transmitter.

amplified in an amplifier with a pass band of 105 to 125 Mc. The output from this amplifier is applied to the Sperry SAC-19 synchrodyne, as in the case of a repeater.

The terminal receiver, shown in Fig. 5, includes the same units as the receiver section of the repeater, except that an rf local oscillator and a power supply must

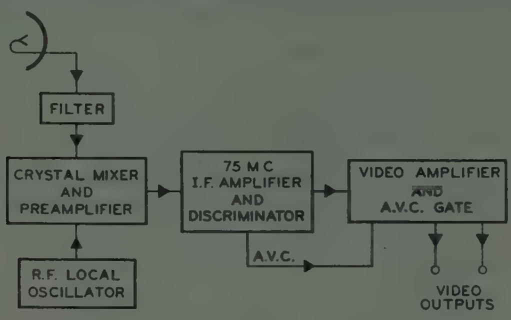

Fig. 5—A terminal receiver.

be added to supply the frequency which, in the repeater, comes from the transmitter local oscillator. However, this oscillator is stabilized in the same manner as the transmitter local oscillator. Adequate frequency stability is achieved by operating the Sperry SAC-19 tube as a precision oscillator. As an oscillator, both cavities are tuned to the desired frequency, and the feedback is arranged through a loop containing a precision cavity having a loaded Q of about 10,000. Since the temperature coefficient of the cavity is very low, in the order of two parts per million per degree centigrade, the oscillator may be depended upon to hold the transmitter frequency well within the FCC tolerance of ± 0.05 per cent, and what is more important, to keep the received signals within the acceptance of the various intermediate-frequency amplifiers throughout the system.

From the foregoing, it can be seen that at a repeater station, if the incoming frequency should for any reason be high or low, the change in the outgoing frequency would be in opposite sense, and of the same magnitude. At the next repeater in the chain, the reverse action would take place. Since the local oscillator and the crystal oscillator in the repeaters are both very stable devices, and since the AFC system of the terminal transmitter holds the synchronizing signal tips at a fixed frequency, the arrangement insures that the signal swing is centered on the receiving discriminator for any reasonable number of repeaters. In the later design, known as the VLR-3, the frequencies have been rearranged to make local oscillator variation at a repeater cancel out. This will also be an improvement because the video polarity will be the same at each repeater.

The sound program portion of a television transmission should also be handled over a radio relay system. To date, no adequate method of multiplexing the two signals on a single television relay channel has been found because of cross-modulation limitations.

The provision of separate radio equipment for the sound channel seems the best approach, as it permits multiplexing of similar material, and provides a method of deriving the much-needed service channel for maintenance of the system. However, many parts may be common, such as antennas, waveguides, local oscillators, power supplies, etc. Obviously, the sound and service channels should be two-path or duplex systems.

Fig. 6 shows the antenna. The waveguide lines are run up the towers and are connected to the antenna with flexible waveguide sections. The antenna reflector is a 4 foot by 8 foot truncated parabola having a focal length of 35.8 inches. The parabolic reflector is mounted with the long dimension vertical and is excited by a vertically polarized horn feed which is held to within an inch of the focal point by four stainless steel rods. Two sus-

Fig. 6—Antenna.

ceptance tuners are mounted on the horn, just behind the mouth. These tuners are adjusted in the laboratory, and then soldered in position. A pressurizing window made of polystyrene, approximately one-half wavelength thick, is used with a rubber gasket to seal the end of the horn. This window is also part of the tuning system. The voltage standing-wave ratio of the horn-and-dish assembly, when measured at the input of the horn feed on the back side of the paraboloid, is under 1.1 throughout the frequency band from 6,115 to 6,235 Mc. These antennas have a power gain of approximately 7,500, a horizontal half-power directivity of 3 degrees, and a vertical half-power directivity of approximatey $1\frac{1}{2}$ degrees. An antenna mounting system, providing both vertical and horizontal adjustments, makes possible the accurate aiming of the paraboloid toward the next repeater station.

RG-50/U, ($1\frac{1}{2}$ inches$\times\frac{3}{4}$ inch$-\frac{1}{16}$ inch wall) silver-plated brass waveguide is used for connecting the antenna systems to the equipments. Silver-plating on the inside of the guides reduces attenuation to the minimum. The entire antenna line is pressurized by a dry-air unit.

Fig. 7 shows the reversing antenna switches which are supplied at the repeaters and terminals to connect the eastbound and westbound antennas to particular transmitters and receivers. These antenna switches are mounted on the top of the cabinets which contain the equipment, and connections are made to the waveguide runs by means of flexible waveguide sections.

The waveguide reversing switch in the antenna system may be operated remotely by the broadcast station operator at the terminal, as the mechanism is is solenoid-operated.

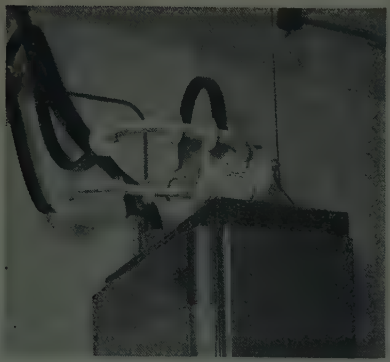

Fig. 7—Waveguide switch.

Fig. 8 shows a front view of two repeaters, one with the front cover panels removed. Metering indications and operating adjustments are accessible with the covers in place. When these covers are removed, all the apparatus is completely accessible. The upper section comprises a relay transmitter, and the smaller section directly below comprises a relay receiver; remaining rack units are power supplies.

The transmission performance of the television relay circuits between New York and Philadelphia is best shown by a series of photographs. For convenience, the

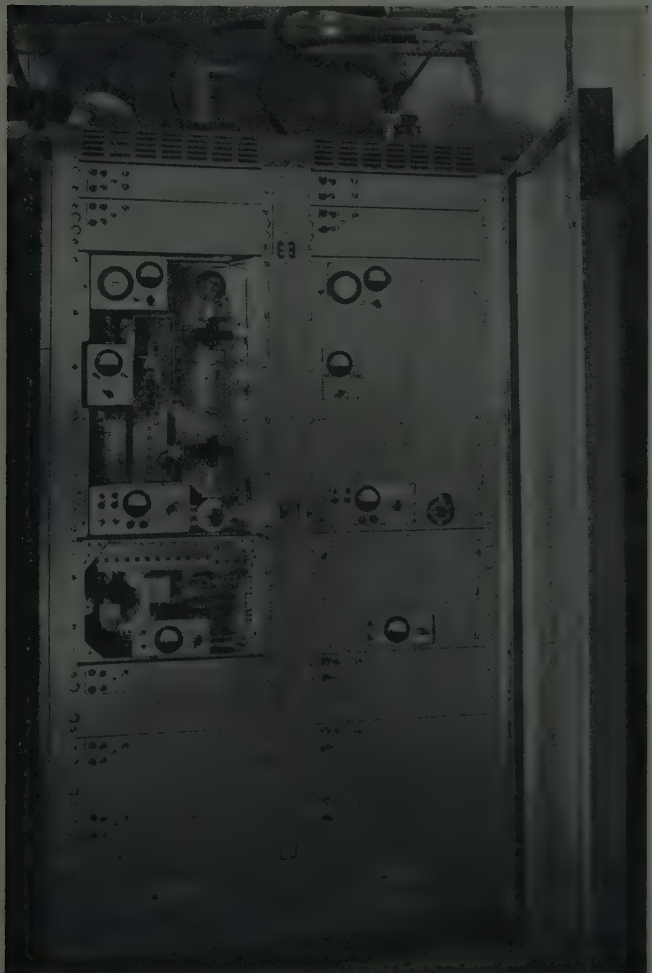

Fig. 8—Front view of two repeaters, one with front cover panels and door removed.

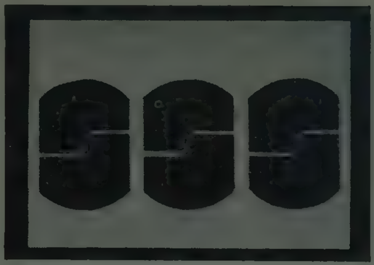

Fig. 9—Transient response of the two New York-Philadelphia television relay circuits connected in tandem.

two circuits were connected together at Philadelphia, which made it possible to impress a signal at New York and photograph this signal before and after transmission over the two circuits in tandem. Fig. 9 is a composite photograph of three oscilloscopic images of a 100-kc square-wave signal. The first image is the output wave of the signal generator. The second image shows the wave after it had been transmitted to Philadelphia over the first circuit and returned to New York over the second circuit. The third image shows the wave after removing the overshoot by transient response adjustments in the video amplifier.

Fig. 10 shows the CBS test pattern photographed on a DuMont monitor after the signal had been transmitted to Philadelphia and back to New York. Within the limits of photographic resolution and the accuracy of the monitor, no distortion or ghosting is apparent.

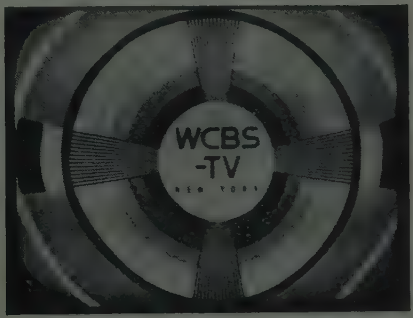

Fig. 10—Test pattern transmission over the two New York-Philadelphia television relay circuits connected in tandem.

The resolution provided by the relay is more than adequate for the transmission of this particular test pattern, and the contrast is well preserved. Microwave relay systems are already playing an important part in Western Union's mechanization and plant improvement program. Present engineering contemplates the sharing of towers and other physical plant along many of these routes by intercity television circuits, should future circumstances make it desirable to furnish such facilities. These circuits may also be used to secure large numbers of voice bands for facsimile message transmission, or larger bands for high-speed facsimile telegraph. Thus it can be seen that microwave radio offers the great promise of providing adequate facilities economically for Western Union's many services, as well as the means for expanding its field of public communications.

The Design of Electronic Equipment Using Subminiature Components*

M. L. MILLER†

Summary—**This article discusses briefly the advantages and disadvantages of subminiature design, and indicates the amount of space that can be saved by such design. It discusses in greater detail the major problems of design, and presents some practical suggestions concerning the methods and materials found most useful.**

For several years there has been an ever-increasing interest in the construction of electronic equipment using subminiature tubes and exceptionally small components. Various laboratories and development groups have collectively constructed nearly every type of applicable circuit using such components. Since the method of design using small components is somewhat different than the regular design procedure, it is felt that a review of some of the more pertinent facts is in order. The information presented here, although probably applicable in part to all subminiature tubes, is based upon experience with high-performance tubes of the heater-cathode type used in circuits where ten or more such tubes are in a single piece of equipment. Many of the data are, of course, applicable to design of compact equipment using any tube type.

I. Advantages of Subminiature Design

THE USE OF subminiature components, in general, is undertaken only because a saving of size and weight or a more rugged construction can be achieved. There are certain applications where such tubes are useful because of their better high-frequency characteristics, but such will not be treated specifically.

1. Space Saving

The use of small lightweight components results directly in a weight savings, but, as will be pointed out later, it is not always possible to save as much space as seems at first possible. Since the amount of space savings achievable depends upon the operating temperature, such will be discussed following the discussion of temperature.

2. Shock Resistance

The susceptibility to damage from shock or vibration of any structure depends, among other things, upon its moment of inertia and the stiffness of the mounting. Thus, the reduction of the mass of the structure improves the ruggedness of components, providing that the stiffness of the mounting does not suffer. Many small components are mounted by their leads. With subminiature components, such leads can be made quite short which further improves the ruggedness of the equipment. In general, the lead stiffness does not reduce in proportion to the mass of the component. For somewhat the same reasons, the components themselves will be more rugged. Thus, improved ruggedness, as well as space and weight savings, can be achieved by simple means, providing subminiature components are used, and it is, in general for these reasons that such components are used.

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II. Disadvantages of Subminiature Design

The first difficulties encountered in using subminiature components are generally not those which prove to be of major importance. Lack of familiarity with compact design generally handicaps the early efforts of any group undertaking such work, with the result that there is generally an overemphasis on such activities as deleting terminal strips, using no chassis, and combining components into exceptionally compact structures to save additional space. The major difficulties connected with such design are, in their approximate order of importance, as follows:

1. High temperature resulting in reduced life and reliability.
2. Difficulty of parts replacement.
3. Circuit difficulties at low frequencies.
4. High cost.

1. High Temperature

The high temperatures are a direct result of reducing the size of the equipment without a proportional reduction of the power input. The effect of this on the design of different types of cirucits will be discussed later.

2. Parts Replacement

The difficulty of parts replacement can be greatly relieved by careful design, but, under the best of circumstances, the compact design generally leaves little room for parts removal, and frequently it is necessary to remove several parts to replace a single one. This difficulty becomes greater because the removal of a part in close quarters frequently damages it, and because most subminiature tubes are soldered into the circuit. Troubleshooting by parts replacement is nearly impossible. These difficulties require that consideration be given to the use of small replaceable subassemblies. If this is not done, greater skill of maintenance personnel is required, and, frequently, special test equipment must be devised for the use of maintenance personnel.

3. Difficulties at Low Frequencies

The difficulty of design for low frequencies stems from the desire to minimize the size of the components. At frequencies below about a kilocycle, transformers or chokes become too bulky to be considered except as a last resort. There is an ever-present urge to use large-value resistors as grid leaks, and thus reduce the size of coupling capacitors in resistance-capacity coupled am-

plifiers. This leads to difficulties with capacitor leakage because of the high ambient temperatures, and requires tubes with exceptionally small grid conductance.

4. High Cost

The high cost of the equipment stems from three sources. The components themselves are generally rather special and thus are expensive. This cost may go down with increased use, but at the present time component costs can be expected to be from two to twenty times as expensive as larger components that would serve the same purpose. The cost of design for most circuits can be expected to be somewhat greater because the final layout must be made by exceptionally skilled personnel after careful planning, and frequently after the construction of several prototype units to study the design. The final cost differential results from the cost of replacing any parts found to be defective or of improper value. To reduce this cost of replacement it will be found economically desirable to increase the expenditures for incoming inspection.

III. Effect of High Temperatures

With octal-base tubes and orthodox construction, it is seldom found necessary to use special means of heat dissipation of electronics equipment. With subminiature components, heat dissipation can become one of the major considerations of design.

There has been set up an empirical relationship between the number of watts to be dissipated and the area of the case capable of dissipating such wattage into still air at room temperature. The range of values of interest are from 6 square inches per watt to $1\frac{1}{2}$ square inches per watt. With reasonable care in construction and component selection, almost any electronic circuit can be made to operate continuously with 6 or more square inches per watt, and almost no piece of electronic equipment can be expected to operate continuously with less than $1\frac{1}{2}$ square inches per watt. Between these two limits the frequency of operation, the band pass, the case material, the thermal bonding between dissipative elements and the case, the required stability of the equipment, and the presence or absence of automatic control or compensation circuits all enter into the maximum workable dissipation per square inch. Fig. 1 shows the relationship between case temperature and square inches of aluminum case per watt.

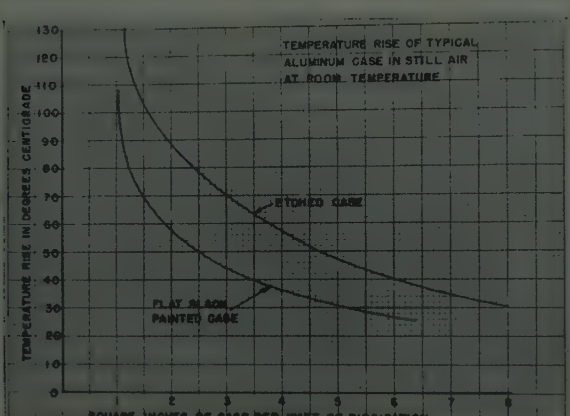

Fig. 1—Plot of temperature rise above ambient of a typical aluminum box having about 150 square inches of surface.

1. Broad-Band High-Frequency Amplifiers

Broad-band high-frequency amplifiers whose gain and frequency stability requirements are average can be built having case areas of 2 to $2\frac{1}{2}$ square inches per watt. Such units operate satisfactorily with case temperatures approaching 100°C, providing components are picked carefully. Such high dissipations per square inch are possible because the circuit Q's are low, the capacitors small, the tuning not exceptionally critical, and most of the dissipation is in the tubes where it is easily conducted directly to the case or outside of the unit. Because such amplifiers must be constructed "in line," it is very difficult to reduce the case area below the figures given. Thus, the maximum dissipation per square inch is unknown, but is probably close to the above.

2. Other Alternating Current Amplifiers

Low-frequency or narrow-band amplifiers whose frequency or gain characteristics are critical must be allotted larger case areas, generally of the order of 3 to 5 square inches per watt. This is true, in part, because paper capacitors are necessary, effective circuit Q's may be high, or narrow tolerances must be held during operation.

3. Direct-Current Amplifiers

High-gain dc amplifiers or low-gain dc amplifiers having exceptional stability requirements may require cases having over 5 square inches per watt.

IV. Size of Equipment

In general, a reasonable measure of the complexity and power consumption of electronic equipment is the number of tubes. Heater-cathode subminiature circuits generally consume on the average of 2 watts per tube.

1. Smallest Size

To estimate the smallest possible size units, it can be assumed that the equipment is to be contained in a long flat box $\frac{1}{2} \times 2$ inches and that there should be a surface of $1\frac{1}{2}$ square inches per watt. The spacing between tube centers would be 0.6 inch resulting in 0.6 cubic inch per tube. This probably is a more compact design than can be achieved with any very useful circuit. About 0.9 cubic inch per tube is the most compact practical design achieved to date. It is a 60-Mc intermediate-frequency amplifier using printed circuits.

2. Larger Units

If the shape of the elongated box is changed to 2 inches by 2 inches in cross section to permit the use of

larger components and a surface of 3 square inches per watt is allowed, the tube centers will be 0.75 inch and the resulting volume will be 3 cubic inches per tube. This is about the minimum size that can be achieved with diversified circuits using separate components.

If twenty tubes consuming 40 watts are to be packaged in a cube so that the area of the container is 120 square inches (3 square inches per watt), the sides of the box will be about 4.5 inches. The volume of the box will be about 90 cubic inches or about $4\frac{1}{2}$ cubic inches per tube.

Even this is a "small" container for a twenty-tube circuit, but it might contain considerable unused space if designed using subminiature components.

From the above it can be seen that size, although dictated by case area, can be reduced to a value far below that possible with larger components.

V. Some Thermal Design Considerations

The amount of space that must be used to provide the necessary case area, and at the same time just enclose all of the parts, is at once the first, and essentially the only, problem in design after a workable schematic diagram is completed. All other design problems only contribute to this solution. However, a knowledge of the required case area, the dimensions of the largest parts, and the desired shape and size of the finished product will generally lead directly to the first trial solution.

1. Thermal Bonding

The first consideration of design must be good thermal bonding between the major heat sources and the case. In most designs it will be found very helpful to bond the chassis to the case as well as possible. The tubes should be contained within close-fitting metal sleeves which are in contact with the chassis over a reasonable area. Whenever possible, the bonding should consist of a metal-to-metal contact of at least $\frac{1}{8}$ square inch of surface per tube. Because of their good thermal conductivity, aluminum or copper is recommended as a material for the sleeves.

2. Chassis Material

Aluminum chassis material is recommended wherever possible. Where such metal is not sufficiently strong in thin sheets, thick aluminum chassis plates ($\frac{1}{8}$ inch thick) are recommended because they offer good stiffness, light weight, good thermal conductivity, and can be tapped, thus eliminating nuts.

3. Case Material

All of the dissipation figures given in III above were based upon experience with equipment dissipating from 15 to 150 watts and having aluminum and/or steel cases. The temperature gradient in the steel sheet may be so large as to reduce the effective case area unless carefully distributed thermal bonds between heat sources and case are provided. Thus, steel cases are somewhat inferior to aluminum cases where there is only a small bonding area between chassis and case.

4. Forced Air Cooling

Where the wattage dissipation per square inch exceeds the values given above, it is necessary to use forced air cooling or some other suitable means of heat dissipation. Where such is feasible, the compactness of the equipment is dictated by the size of the components and the space required to ventilate them.

VI. Selection of Components

The selection of components for use in subminiature electronic design consists, in general, of finding small components that will stand the high temperatures.

1. Tubes

Considerable has been written about subminiature tubes and their application. There are now available tubes of almost every type necessary to design any electronic circuit, excepting those which require tubes with large plate dissipations. For the majority of applications tube size is of secondary importance, providing the tube is in a T-3 (or smaller) envelope. If it were possible to reduce the power consumption of the tube in proportion to its size, then any reduction of size and dissipation would be advantageous.

2. Resistors

Standard half-watt insulated carbon resistors will supply substantially all resistor needs. The insulation is important, and the use of narrow tolerance parts will be found advantageous because somewhat larger changes (of all parts) due to temperature changes can be expected than with less compact equipment. Other resistor requirements can also be met with standard parts. Small variable resistors are available, but there remains a need for a subminiature locking-type potentiometer. For units that are to operate with case temperatures approaching 100°C, it may be necessary to use resistors at less than their full rating. The manufacturers' recommendations should be followed in this regard.

3. Capacitors

Electronic circuit designers have long considered a capacitor as a device drawing only reactive current and serving as an open circuit for dc. At temperatures above 100°C, this idea must be modified greatly unless capacitors of exceptional physical size or small capacitance are used. For capacitors of the size range of 0.01 to 1 μf operating at 150°C, a dissipation factor of 10 per cent is above average. In this size range, the dc leakage at 150°C is less than 10 megohm-microfarads for any known capacitor suitable for use in subminiature equipment. Most manufacturers of small capacitors do not recommend that they be operated above about 100°C and promise very little for their performance at higher temperatures. Some of the lowest leakage capacitors found to date are the new metalized paper capacitors, but these generally have leakages of about 1 megohm-microfarad at 150°C. Molded mica and ceramic capaci-

tors are generally not much superior as far as leakage is concerned, probably because of case leakage. A few samples of uncased mica capacitors showed leakages of nearly 10 megohm-microfarads at 150°C. In general, no satisfactorily high leakage resistance capacitors are available for high temperatures. The circuits must be so designed as to permit considerable capacitor leakage.

Electrolytic capacitors will operate for a considerable period at 100°C or slightly above.

Variable capacitors are generally to be avoided whenever possible because of their size. However, compression mica trimmers generally are satisfactory for circuits that are not excessively critical. Cycling in temperature a few times sometimes improves the stability of such trimmers.

4. *Inductances*

Small high-quality inductances are available from many component manufacturers. Although such units are generally not recommended for use at high temperatures, no difficulty has been encountered with them in subminiature equipment. No undue reduction of life has been noted, and the increase of the copper loss at high temperatures is generally not sufficient to cause difficulty. Wherever the shielding can be dispensed with, iron core units should be selected without cases to save space.

5. *Transformers*

The size of iron-core audio transformers is dictated by the same considerations as inductances. Thus, the above applies to them. With power transformers it is possible, by use of silicon and glass insulation, to operate the transformers at high flux and current densities with remarkable savings of space and weight. Such operation is permissible only if the life requirement of the unit is reduced, because the interturn insulation must remain the ordinary wire enamel to save space. An example of such a unit is a 350 VA, 11-winding, 3-phase, 400-cycle power transformer that occupies 15 cubic inches and weighs 21 ounces.

6. *Miscellaneous*

Other components sometimes require greater care in their selection than the major ones, but space does not justify detailed discussions of such equipment. Briefly, wire must be the highest quality available (glass-insulated wire is nearly essential); the smallest relays available are quite suitable for most applications, although melted wax from other components can cause fouling of the contacts; terminal lugs and mounting strips are generally too large, but specially constructed ones are excessively expensive. Suitable small connectors are difficult to find. Doubtless there will be available at some time in the future a connector having most of the good points of the many now proposed or available.

VII. Specific Circuit Design

In general, the use of subminiature components does not alter the circuit design but creates a new emphasis upon certain circuit problems. The major design problems generally are mechanical. However, certain electrical considerations are of importance.

1. *Radio-Frequency and Intermediate-Frequency Design*

After the smallest available components have been found, the only design problem is that of providing the increased shielding required because of the reduction of the size of the amplifiers. Exceptional care in the placement of the components must be used.

2. *Audio and Video Design*

Because the size of transformers and inductors is not subject to much reduction, selective-frequency circuits in the low audio band require the use of filters or tuned circuits using only resistors and capacitors. The capabilities and limitations of *R-C* filters (including twin-tee filters) are well documented. Therefore, it is only necessary to adapt such circuits to the problems at hand.

The *R-C* type of filter generally requires greater gains in the associated amplifiers, but even the size of an additional gain stage is small in comparison to the size of an inductor.

For broad-band audio or video amplifiers resistance-capacity interstage coupling is essential, and the problems of capacitor leakage become important. The cathode follower will be found to be the standard impedance-matching device in subminiature design. In general, transformers will be found useful as a last resort.

Power at audio or video frequencies is limited by the plate dissipation of subminiature tubes with the result that one or two watts is the upper limit. At any frequency, the number of subminiature tubes that can be paralleled successfully is an interesting engineering problem. Because the tubes are soldered into the circuit, it is frequently difficult to find a defective tube, and it is sometimes difficult to remove the tube without breaking the leads on one or more other tubes or parts. The result is that the cost of finding and replacing a single defective tube may be quite large in comparison to the cost of replacement of a miniature tube which would have had greater plate dissipation and occupied at most only a small additional space.

3. *Direct-Current Amplifiers*

Only the warm-up characteristics of subminiature dc amplifiers can be expected to differ from ones having larger components. Since the amplifier will operate at higher temperature, special care must be used in selecting and placing components to minimize the temperature drift of the amplifier. If this is done, and special care is used to insure good thermal bonding, the warm-up characteristics of the subminiature amplifier may be superior to one of orthodox design. The good thermal bonding will result in rapid approach of all components to the stability temperature, and thus shorten the warm-up period. This improved warm-up characteristic can be achieved at any frequency by careful design.

4. General

With tubes that are soldered into the circuit it is feasible to adjust some circuit parameters by parts replacement. Since the operation of tube replacement requires the use of a soldering iron, if the necessary parameter adjustment can be made by substitution of one or two standard RMA resistors, most of the space required for a potentiometer or rheostat can be saved by simple parts replacement. Such adjustments require more time, but, if the layout is so made that the required part is easily accessible, the time used for adjustment can be made short in comparison to that of tube replacement.

VIII. Layout

Various laboratories have produced differing solutions to the problem of layout of subminiature equipment. Each of these solutions is based upon the particular requirements of the laboratory making the equipment. At the present time there is no means of passing comparative judgment upon the various solutions. A few of the important considerations will be outlined here.

1. Layout Using Components

To save the maximum amount of space it is necessary to group the tubes into clusters of three or four so that common terminals can be used as much as possible. This, of course, applies only to the frequency bands where parallel (or series) operation of heaters and close proximity of parts are permissible. As previously mentioned, the tubes should be supported in the clusters by metal sleeves which offer good thermal bonding to the chassis. Fig. 2 shows tube clusters before mounting. Terminal lugs can be arranged in a rectangle around the tube clusters and the parts supported on the lugs with the shortest possible leads. With this type of construction, capacitors can generally be placed so that they are not adjacent to resistors or tubes having high dissipation, which results in less capacitor leakage.

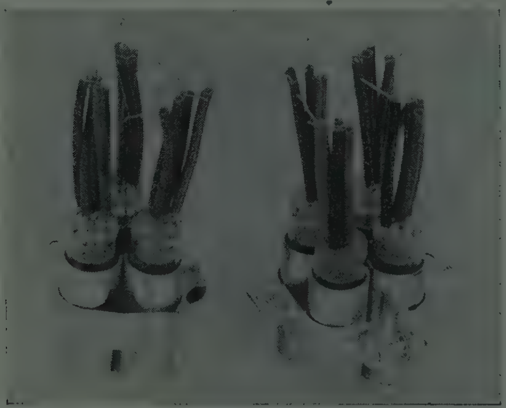

Fig. 2—Typical tube clusters for T3 tubes ($\frac{3}{8}$-inch diameter) used in compact equipment to take advantage of common terminals to save space.

The most compact design generally requires that components and/or wiring appear on both sides of a chassis while the tubes protrude through the chassis. By orienting tubes properly, it is possible to have wiring on both sides of the chassis, but it is frequently advantageous to mount potentiometers, chokes, trimmers, and transformers with pigtail leads on one side of the chassis between tube clusters while retaining the major portion of the wiring all on one side. Fig. 3 shows an if strip using this type of construction.

Fig. 3—Typical 5-stage intermediate-frequency amplifier using subminiature components. This amplifier chassis is $6\frac{3}{4}$ inches long by $1\frac{1}{8}$ inches wide.

2. Printed Circuits

Printed circuits appear as a final means of further compacting equipment where the space saved by such means can be used. The space limitations set up by the heat generated are generally of such a nature that more orthodox components can be used. In general, the cost of printed circuits cannot be justified for development or small production work unless full advantage can be taken of the space saved. At least one laboratory has demonstrated the remarkable space savings that can be achieved by using printed circuits on high-gain intermediate-frequency amplifiers where nearly all components can be printed on glass or ceramic, and thus withstand exceptionally high temperatures. Reference should be made to National Bureau of Standards Circular 468 and Miscellaneous Publication 192.

IX. Potting and Casting

Where it is desired to achieve the ultimate in ruggedness, to prevent corona at high altitudes, or minimize the effects of humidity, complete units can be potted in wax or cast into plastic. The potting wax is disadvantageous because most subminiature equipment operates near the melting temperature of suitable waxes.

1. Casting

The casting materials, such as polyester-styrene resins, set at room temperature and will withstand high temperatures after setting. Electronic equipment can be cast into such material with the result that the cast unit will withstand more vibration or shock than the tubes contained within it. Of course, moving parts must be excluded from the cast sections.

2. Size and Weight of Cast Units

Once cast into such material, there is no possibility of parts replacement. Thus, such cast units must themselves be considered as replaceable units. This limits the number of parts that can be cast into a single unit, and suitable means for determining the number of parts based upon their reliability must be devised by the design group. The problem of making suitable interconnection between such cast units requires either small plugs or the use of cast-in soldering terminals and/or pigtails.

Cast units are generally somewhat heavier than the equivalent units assembled on an aluminum chassis.

3. Typical Construction

One of the best designs found to date is to assemble all parts petween two thin sheets of insulating material in such a manner that the leads protrude through small holes in the material. The connections are made by bending and soldering the leads outside the insulating sheets. With some care in design the resulting assemblage will be sufficiently rigid to test before casting. After test, the unit is supported in a mold and the casting resin poured over it and allowed to set. For most applications the casting resin will function as both support and a heat-conducting medium; thus, no chassis is required. By casting mounting lugs or screws into the assembly it can be fastened to other parts in any manner desired. Fig. 4 illustrates this type of construction before the casting resin is poured.

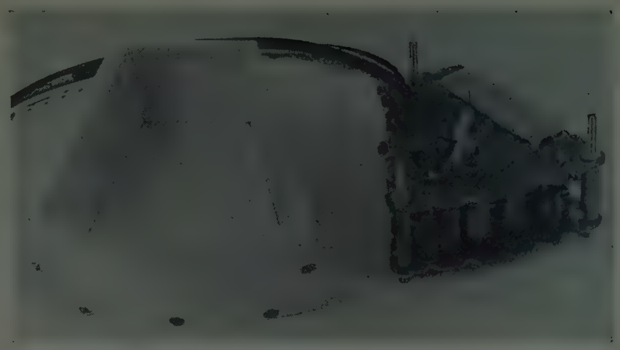

Fig. 4—Subminiature assembly ready to be cast in polyester-styrene resin, and the mold in which it is to be cast.

X. Conclusions

The serious application of subminiature components to electronic design is rather recent with the result that a universally applicable layout and design procedure has not yet been found. It is probable that no single solution will satisfy all needs. At the present time, it is hoped to offer only sufficient general information to persons starting new designs so that they will be able to judge the suitability of subminiature design for their specific problems, or so that once started, they will not repeat all the errors of the earlier workers in the field. It is hoped that the information contained in this discussion will provide such minimum guidance.

Electron Beams in Axially Symmetrical Electric and Magnetic Fields*

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Summary—The problem considered in this paper is the formulation of the equations governing the motion of an electron beam in axially symmetrical magnetic and electric fields. The equations are obtained for the trajectories of the electrons along the outer edge of the beam for the most general case, in which there are both axial and radial components of the fields. It is shown that, as a result of symmetry, the combined effects of the electric and the magnetic fields can be expressed as a single generalized potential function which depends only on the axial and radial space co-ordinates. This permits one to express the axial and radial force components as the axial and radial components of the gradient of this potential function.

Numerical solutions have been obtained by numerical integration for the trajectories in a uniform magnetic field. Curves are presented in normalized form, giving the results of these solutions for cases likely to be encountered in practice. It is shown that there exists an equilibrium radius for which the net radial forces acting on the electrons is zero, and that the outer radius of the beam will oscillate about this equilibrium value, the amplitude being nonsymmetrical and depending upon the initial conditions, and the wavelength (distance between successive maxima) depending upon the amplitude.

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Introduction

FOR CERTAIN TYPES of electron tubes, it is desirable to keep an electron beam to a small diameter for a considerable length. It is not very difficult to do this if the electron beam density involved is not large, as in a cathode-ray tube or electron microscope, in which the repulsion forces between electrons are negligible. As a matter of fact, the electron optics used in such kinds of tubes treat the trajectories of individual electrons as if the other electrons do not exist. Usually pure electrostatic means of control are sufficient to control the motion of an electron beam as one desires.

In high-power klystrons[1] or traveling-wave tubes,[2] the electron beam current required is relatively large. At the same time, the beam has to be kept small to get better beam-field interaction of considerable length to yield good efficiency and gain. For obvious reasons such beams are confined within a unipotential cylinder or conductors, so that no electrostatic field can be used to counteract the space charge repulsion. It is necessary in such cases that some extra forces be provided, in order to accomplish the above objectives in electron beam requirements.

It has been known that an electron beam has the tendency of following magnetic lines of force. Therefore, one obvious way of obtaining a long beam of small diameter is to immerse the beam along the magnetic lines of force of a longitudinal magnetic field all the way down to the cathode region where the beam originates. As the space-charge repulsion force causes the electron to cross the magnetic flux, the resultant action of the magnetic field is always in such a direction as to bring the electron back to the same magnetic line of force which it originally followed. The stronger the magnetic field is, the less distance the electron has to go astray before it comes back. Thus by increasing the magnetic field strength indefinitely, one can make the undulations as small as possible.

However, in a practical tube, if very high current density is desired, the magnetic field required to make a beam with tolerable undulations may become prohibitive. Actually, this brute force method of using a strong magnetic field with completely immersed cathode has been adopted generally in many designs of the so-called "magnetically focused tubes."

The present paper considers the general equations governing the motion of an axially symmetric electron beam in an axially symmetric magnetic field, in order to determine the magnetic forces involved under different circumstances. From this study, it becomes evident that instead of letting the electron beam follow the magnetic flux closely from the beginning to end, the proper way is to deliberately let the beam cross the magnetic flux lines in a controlled region before the beam proceeds to follow the magnetic lines of force. By so doing, one can completely eliminate the undulations and obtain an indefinitely long parallel beam as long as the magnetic field exists. And, it is no longer necessary to use as strong a magnetic field as one can obtain. For a given system, the magnetic field required to keep a given size parallel beam is a definite quantity. The least amount of magnetic field is required when the magnetic field is completely excluded from the cathode region. The following equations are developed to cover all the cases with various amounts of magnetic flux included in the cathode region. Numerical integrations are carried out so that the beam trajectories can be easily calculated if the magnetic and electric field configurations of the system are given.

Mathematical Statement of the Problem

Problems of electron dynamics[3] can be treated in terms of the conservation of energy considerations, or more specifically in terms of a Hamiltonian function

$$H = \tfrac{1}{2}m(\dot{r}^2 + \dot{z}^2 + r^2\dot{\theta}^2) - e\phi = 0, \tag{1}$$

where H is the algebraic sum of the kinetic energy expressed in cylindrical co-ordinates of r, z, and θ, and the potential energy, ϕ. As written, this equation assumes that the electrons all originate from a common cathode at zero potential and with zero initial velocity. We will find it convenient to write the Hamiltonian in terms of generalized momenta, p_r, p_z, and p_ϕ, and use the three Hamiltonian equations

$$\frac{dp_r}{dt} = \frac{\partial H}{\partial r} \tag{2}$$

$$\frac{dp_z}{dt} = \frac{\partial H}{\partial z} \tag{3}$$

$$\frac{dp_\theta}{dt} = \frac{\partial H}{\partial \theta}. \tag{4}$$

As a result of our assumed axial symmetry, the external magnetic field can be expressed in terms of a single component of the magnetic vector potential, A_θ, where the magnetic field is given by the curl of the magnetic vector potential, in this case.

$$B_z = \frac{1}{r}\frac{\partial rA_\theta}{\partial r} \tag{5}$$

$$B_r = -\frac{\partial A_\theta}{\partial z}. \tag{6}$$

In performing the partial differentiations in equations (2) to (4), the Hamiltonian must first be expressed explicitly in the generalized momenta and space coordinates. The velocities are eliminated from the Hamiltonian through the following substitution.

$$p_r = m\dot{r} \tag{7}$$

$$p_z = m\dot{z} \tag{8}$$

$$p_\theta = mr^2\dot{\theta} - erA_\theta, \tag{9}$$

and the Hamiltonian can then be written

$$H = \frac{1}{2m}\left[p_r^2 + p_z^2 + \left(\frac{p_\theta}{r} + eA_\theta\right)^2\right] - e\phi = 0. \tag{10}$$

Starting with these equations, we will show that the angular velocity $\dot{\theta}$ must be a function of the space coordinates only. As a result of this fact, we can ignore

[1] E. R. Warnecke, P. Guerrard, and C. Fauve, "Sur les Effects de Charge D'espace les tubes a modulation de vitesse a groupement par glissement," *Ann. de Radioelectricté*, Tome II, no. 9, pp. 1–8; July, 1947.

[2] J. R. Pierce, "Theory of the beam-type traveling-wave tube," Proc. I.R.E., vol. 35, pp. 111–123; February, 1947.

[3] J. C. Slater and N. H. Frank, "Introduction to Theoretical Physics," McGraw-Hill Book Co., New York, N. Y., 1st Ed.; 1933.

the θ co-ordinate of the electron motion and study the electron trajectories in the r, z plane. In this plane the motion of the electrons is subject to a potential function, which is the difference between the ordinary electrostatic potential and an additional function equal to the rotational kinetic energy expressed as a function of the space coordinates. This potential function is

$$e\phi - \frac{m}{2} r^2\dot{\theta}^2.$$

The next step in the analysis consists in solving for the resulting electron trajectories. This has been done both by numerical integration and by constructing a mechanical model of the potential function on which small balls, representing electrons, may be rolled. Complete solutions are given for the constant field case with various amounts of magnetic field threading through the cathode. A method of treating slowly varying magnetic fields is also proposed, and an extension made to the case of more rapidly varying fields by treating the problem in steps.

Electrostatic and Magnetic Potential Functions

Before proceeding with the detailed solution, we will have to inquire into the meaning of the potential functions that we will find convenient to use.

Let us consider an actual beam system as shown in Fig. 1(a). In general, electrons from a cathode shown as being at a zero potential are accelerated and focused by a series of apertured electrodes (A_1, A_2, and A_3) perhaps with an auxiliary magnetic field produced by coils B_1 and B_2. The path of an edge electron (in the r, z plane) is shown as curves OPQ, while an interior electron might follow some other path $O'P'Q'$. The curves OPQ can be thought of as being either the envelope of the beam or as being the trajectory of an edge electron referred to a rotating system of coordinates which follows the rotation of the electron beam. The working portion of the beam will usually be restricted to the region confined by the electrode A_3, where the electrons are traveling at their maximum velocity, and we will concern ourselves mainly with this region.

The total electrostatic potential at any point can be obtained by solving Poisson's equation subject to the prescribed boundary conditions (usually given by the potentials of the confining electrodes). This potential depends upon the magnitude and distribution of the space charge. We will find it convenient to consider this total potential as a superimposition of two potential functions ϕ_L and ϕ_s, where ϕ_L is the Laplacian potential which is independent of the space charge and determined only by the boundary conditions, that is

$$\nabla^2\phi_L = 0 \tag{11}$$

$$(\phi_L)_{\text{electrodes}} = \text{applied potentials.} \tag{12}$$

The second component ϕ_s is the potential satisfying Poisson's equation when all of the electrodes are at zero potentials, that is

$$\nabla^2\phi_s = \rho/\epsilon \tag{13}$$

$$(\phi_s)_{\text{electrodes}} = 0. \tag{14}$$

Fig. 1(b) is a plot of the way in which ϕ_L and ϕ_s on the beam axis might vary for the structure shown in Fig. 1(a). In general, ϕ_s is small compared to ϕ_L in the high-voltage region, a fact that makes the solution of the variable field case much easier.

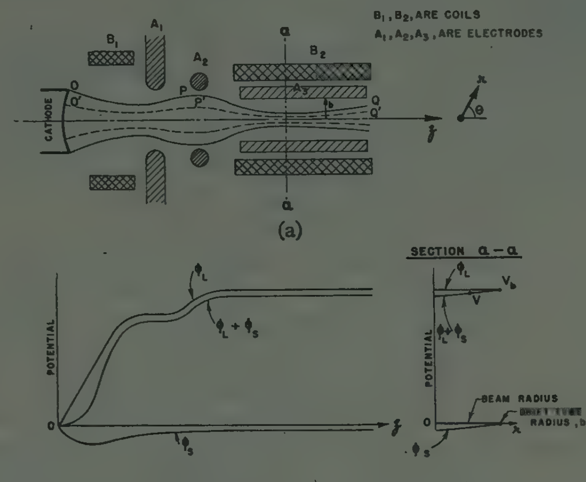


Fig. 1—(a) Schematic diagram illustrating a circularly symmetrical beam system. OPQ is the boundary of the beam and $O'P'Q'$ is the trajectory of an inner electron. A_1, A_2, A_3 are electrodes, and B_1, B_2 are magnetizing coils. (b) Diagram illustrating the relative magnitude of the components of the electrostatic potentials along the axis. ϕ_L is the Laplacian potential, which is independent of the space charge, and ϕ_s is the space-charge potential. (c) A similar plot of the potential functions across a section perpendicular to the axis.

We need to know the value of ϕ_s over the interior of the beam, in order to obtain the axial component of velocity and hence the total charge per unit length at any point along the beam. We will define the ratio of this total charge to the apparent charge per unit length calculated in terms of the total current and potential (V_b) of the confining electrode, that is

$$\xi = \frac{2\pi \int_0^{r_1} \rho r dr}{I \Big/ \sqrt{\frac{2e}{m} V_b}}, \tag{15}$$

where r_1 is the outside radius of the beam, ρ is the space charge density, and I is the total current. The quantity ξ is always greater than unity because of two effects which are additive. In the first place, the z component of the velocity is only one component of the total velocity. Also, the actual potential will sag in the interior of the beam because of space charge (ϕ_s is always negative). In trajectory studies, the quantity ξ can be treated as a nearly constant quantity if the nearby electrode potential V_b is constant.

The space-charge component of the radial electric field for a uniform axial charge distribution is obtainable from Gauss' theorem by dividing the total charge

by $2\pi\epsilon$ times the radius. By integrating this, the value of ϕ_s external to the beam is found to be

$$\phi_s = -\frac{I}{2\pi\epsilon\sqrt{\frac{2eV_b}{m}}}\log_e\frac{b}{r}, \tag{16}$$

where b is the radius of the external electrode. The assumptions underlying (16) must be stated clearly. This equation is true, irrespective of the shape of the beam or of the confining electrodes (of course always limited to axially symmetric systems), if only the total charge per unit length in the beam is constant along the beam. Actually, this equation may be used for most practical cases with a fair degree of accuracy, even though the charge distribution is not strictly uniform.

Inside the beam the values of ϕ_s will depend upon the radial charge distribution. Samuel[4] has shown that a uniform charge distribution can be realized for the case with magnetic field excluded from the cathode. Such a distribution seems not unreasonable for the more general case. The deviation from uniformity occurs only for very high perveance beams, and can be considered as a second-order effect in beams we are likely to deal with in practice. If the charge distribution is approximately uniform, the space-charge component of radial electric field will be proportional to the radius, with the coefficient of proportionality such as to match the external conditions as specified by (16) at the edge of the beam. It can also be shown in the following analysis, with certain restrictions, that the magnetic force is also proportional to the radius. Since the resultant forces are proportional to the radius, we are able to assume that the paths of the interior electrons are similar to those of the edge electrons (that is, having the same axial component of velocity and having the radial component of velocity and displacement proportional to the radius). An interior path in the r, z plane might then be as shown in Fig. 1(a) by the path $O'P'Q'$ where the edge electron path is OPQ. Under this assumption, it is only necessary for us to study the trajectory of the edge electron.

Equation (16) can now be interpreted as defining space charge potential experienced by an edge electron. This potential can be considered to be a fixed function of the space co-ordinates only. It defines the space-charge potential that an edge electron would assume were the beam of such a shape as to permit the edge electron to reach the co-ordinates in question. Since ϕ_L must be a fixed space function, the total potential which should be used in the Hamiltonian (1) when it is applied to the edge electrons is then

$$V = \phi_L - \frac{\xi I}{2\pi\epsilon\sqrt{\frac{2eV_b}{m}}}\log_e\frac{b}{r}, \tag{16a}$$

[4] A. L. Samuel, "On the theory of axially symmetric electron beams in an axial magnetic field," Proc. I.R.E., vol. 37, pp. 1252–1258; November, 1949.

where the V's are used in place of ϕ's to refer to edge electrons only, and r is now used without a subscript to mean the radius of the beam. If numerical values are introduced, this can be written as

$$V = V_b\left[\frac{\phi_L}{V_b} - 0.01515\,\overline{K}\log_e\left(\frac{b}{r}\right)^2\right] \tag{16b}$$

where $\overline{K}$ is the effective perveance in microamperes per volts$^{3/2}$ that is

$$\overline{K} = \frac{\xi I}{V_b^{3/2}} \times 10^6. \tag{16c}$$

We must now give some consideration to the magnetic potential function. We have seen that the applied magnetic field can be expressed in terms of a single component A_θ of the magnetic vector potential. Integrating (5), we can write

$$rA_\theta = \int_0^r rB_z dr. \tag{17}$$

Equation (17) defines rA_θ as a quantity proportional to the magnetic flux enclosed in a circle of radius r. The choice of zero as the lower limit of integration is equivalent to setting A_θ equal to zero on the axis, a choice we are at liberty to make.

We note further that the third Hamiltonian equation, equation (4), may be set equal to zero in view of our assumption of axial symmetry. On substituting the value of p_θ from (9) we have

$$\frac{d}{dt}(mr^2\dot\theta - erA_\theta) = 0 \tag{18}$$

or

$$mr^2\dot\theta - erA_\theta = -er_cA_{\theta c}, \tag{19}$$

where $r_cA_{\theta c}$ is a measure of the flux at the cathode passing through a circle of radius r_c. Equation (19) implies that the electrons leave the cathode with zero angular velocity. Equations (18) and (19) imply that the angular momentum $mr^2\dot\theta$ must change along the electron path in such a way as to keep the generalized angular momentum p_θ a constant. Furthermore, a difference between the angular momentum at any two points at different radial distances can be expressed as the difference between the enclosed magnetic flux for two circles of the specified radii (this is in view of the interpretation of rA_θ noted above). A convenient way to write (19) for the edge electron is to define a radius function g, as shown in Fig. 2, by the radial co-ordinates of a line of flux which goes through the edge of the cathode and write

$$mr^2\dot\theta = e(rA_\theta - r_cA_{\theta c}) = e\int_g^r rB_z dr. \tag{20}$$

This equation shows that the angular momentum is a space function only and is independent of the path of the edge electrons, since the magnetic field is assumed to

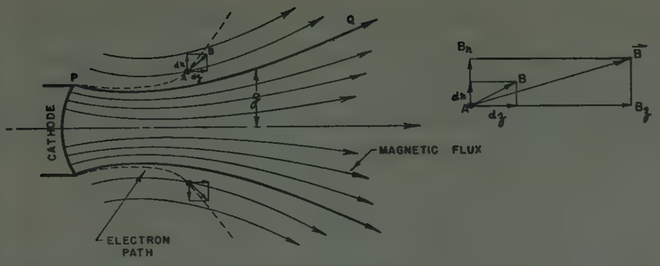


Fig. 2—Diagram illustrating the motion of an electron in a magnetic field. Force acting on an electron in the circumferential direction is proportional to the cross product of the magnetic field $\overline{B}$ and the velocity vector $\overline{AB}$ in the r-z plane. g is the radius of the line of force which passes through the edge of the cathode.

be fixed in space. By way of contrast, $\dot{r}$ and $\dot{z}$ are not space functions, since within the limitations imposed by (1), $\dot{r}$ and $\dot{z}$ may assume any combination of values and are not fixed by the co-ordinates.

Derivation of Trajectory Equations

We are now in a position to specialize the Hamiltonian equations to apply to the edge electrons only and derive the trajectory equation. It follows from (1) that

$$\frac{dH}{dt} = 0, \quad \text{and} \quad H = 0. \tag{21}$$

Differentiating (1), we replace ϕ by V to signify that we are considering an edge electron and observing Euler's convention of differentiation following the motion of a particle where

$$\frac{d}{dt} = \sum_1^3 \dot{x}_k \frac{\partial}{\partial x_k} = \dot{r}\frac{\partial}{\partial r} + \dot{z}\frac{\partial}{\partial z}, \tag{22}$$

since $\partial/\partial\theta = 0$ because of symmetry. Equation (22) applies only to the space functions, which in this case include

$$eV - \tfrac{1}{2}mr^2\dot{\theta}^2$$

as shown in the previous section. Then

$$\frac{m}{2}\frac{d}{dt}(\dot{r}^2 + \dot{z}^2) = \frac{d}{dt}\left(eV - \frac{m}{2}r^2\dot{\theta}^2\right) \tag{23}$$

or

$$m(\ddot{r}\dot{r} + \ddot{z}\dot{z}) = \dot{r}\frac{\partial}{\partial r}\left(eV - \frac{m}{2}r^2\dot{\theta}^2\right) + \dot{z}\frac{\partial}{\partial r}\left(eV - \frac{m}{2}r^2\dot{\theta}^2\right). \tag{24}$$

The operation indicated on the right can be performed only when $\dot{\theta}^2$ is expressed explicitly in terms of space functions.

As a result of the presence of the magnetic field, we see that there exists a new potential function which is given by the local electric potential minus a term which accounts for the rotational energy. The kinetic energy, which must be divided between radial and axial components, is given by the product of this new potential function and the electronic charge. Since $\dot{r}$ and $\dot{z}$ are not space functions, the coefficients of $\dot{r}$ and $\dot{z}$ in (24) must be individually equal to zero. That is,

$$m\ddot{r} = \frac{\partial}{\partial r}\left(eV - \frac{m}{2}r^2\dot{\theta}^2\right) \tag{25}$$

and

$$m\ddot{z} = \frac{\partial}{\partial z}\left(eV - \frac{m}{2}r^2\dot{\theta}^2\right). \tag{26}$$

Equations (25) and (26) are the equations for the trajectory of the edge electrons.

Introducing the value of $\dot{\theta}$ from (19) and replacing A_θ by its equivalent in terms of the magnetic field, we can write

$$m\ddot{r} = e\frac{\partial V}{\partial r} - er\dot{\theta}B_z + mr\dot{\theta}^2 \tag{27}$$

and

$$m\ddot{z} = e\frac{\partial V}{\partial z} + er\dot{\theta}B_r, \tag{28}$$

where the equations are written in such a form as to separate the electric forces, the Lorentz forces, and the centrifugal force. Actually, (27) and (28) can be derived directly from the Hamiltonian differential equations (2), (3), and (4). In the present treatment, the potential function method is preferred, since only a single generalized potential function is involved which can be studied and visualized in terms of physical space models.

Net Radial Force Associated with an Angular Rotation of the Beam

The primary purpose of introducing a magnetic field is to provide some extra force to oppose the space-charge repulsion forces. We should, therefore, examine more carefully the radial forces that result from an angular rotation. As we can see from (27), this force is composed of two terms. One term, $er\dot{\theta}B_z$, the Lorentz force, depends upon the direction of rotation, while the second, $mr\dot{\theta}^2$, the centrifugal force, is independent of the direction of rotation. Fig. 3 is a plot of the net radial rotational force in normalized angular velocity. The two terms of interest are

$$F = -er\dot{\theta}B_z + mr\dot{\theta}^2. \tag{29}$$

If we introduce the notation

$$\omega_H = \frac{1}{2}\frac{e}{m}B_z, \tag{30}$$

where ω_H is the usual Larmor frequency, we can write

$$\frac{F}{mr\omega_H^2} = -2\frac{\dot{\theta}}{\omega_H} + \left(\frac{\dot{\theta}}{\omega_H}\right)^2. \tag{31}$$

To counteract the space-charge repulsive forces, $F/mr\omega_H^2$ in (31) should be negative. In a given mag-

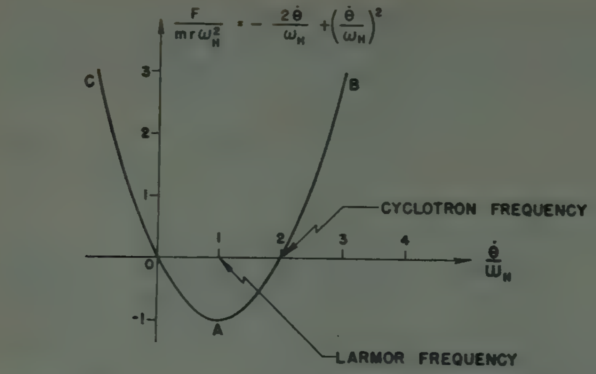


Fig. 3—Net normalized force $F/mr\omega_H^2$ in the radial direction attributed to the rotational velocity $\dot\theta$ of a charged particle in a magnetic field of Larmor frequency ω_H, plotted as a function of $\dot\theta/\omega_H$.

netic field, that is, for a fixed value of ω_H, this force F has a maximum negative value when $\dot\theta = \omega_H$. Conversely speaking, this is also the condition to produce a given force with a minimum magnetic field.

Restricting ourselves for the moment to the uniform field case, that is, with B_z and g both constant, we can integrate (20), obtaining

$$mr^2\dot\theta = \tfrac{1}{2}eB_z(r^2 - g^2) \tag{32}$$

or, in terms of ω_H,

$$\dot\theta = \omega_H\left(1 - \left(\frac{g}{r}\right)^2\right). \tag{33}$$

We observe that $\dot\theta$ is equal to ω_H for the special case where none of the flux threads through the cathode; that is, when $g=0$. For the more general case when $g \neq 0$, $\dot\theta$ depends upon the radius. When g is real and r is less than g, $\dot\theta$ is negative and F is positive (region 0 to C on Fig. 3). When r is greater than g, $\dot\theta$ is positive and less than ω_H (region 0 to A on Fig. 3). When g is imaginary, corresponding to the condition where the flux threading through the cathode is in the opposite direction to that in the region under consideration, $\dot\theta$ is positive and greater than ω_H (region AB on Fig. 3).

The most economical design as far as the required magnetic field is concerned is evidently achieved when none of the magnetic flux threads through the cathode. Under these conditions, the electrons on entering the field will automatically rotate at the Larmor frequency.

If the required magnetic field strength is not of primary concern, the same force F can be obtained with a smaller value of $\dot\theta$ by allowing some of the magnetic flux to thread through the cathode region. This means that more magnetic field is required but that a smaller fraction of the total energy has to be put into rotational form.

For the third possibility, where the flux threading the cathode is in the opposite direction, both the required magnetic field and the rotational energy are increased. The region of the curve in Fig. 3 from A to B is perhaps of academic interest only, and will not be discussed further.

Trajectory Solutions for the Uniform Field Case

We will first consider the case where the magnetic field and the Laplacian electric field are both uniform and where ϕ_L is equal to V_b, the potential of a confining conductor. Equation (16b) reduces to

$$V = V_b\left[1 - 0.01515\,\bar{K}\log_e\left(\frac{b}{r}\right)^2\right]. \tag{34}$$

Equation (33) applies as written, but g is now a constant. In fact, the generalized potential function $(eV - (m/2)r^2\dot\theta^2)$ is independent of z, equation (26) is equal to zero, and the axial velocity $\dot z$ must be a constant. A physical model of the potential function then reduced to a simple cylindrical structure, as will be discussed later.

We can now substitute the value of V from (34) and the value of $\dot\theta$ from (33) into (25) to obtain the general trajectory equation. Our work will be simplified considerably if we introduce a normalizing constant, a, which has the dimensions of a length and has a physical significance in terms of a certain radius which will be explained later. The constant a is defined by the relation

$$a = \sqrt{\frac{1}{2\pi\epsilon}}\left(\frac{2m}{e}\right)^{3/4}\frac{(\xi I)^{1/2}}{V_b^{1/4}B_z} \tag{35}$$

or, in numerical terms (mks units),

$$a = \frac{\sqrt{\xi I}}{1.203V_b^{1/4}B_z} = \frac{\sqrt{\bar{K}V_b}}{1.203\times 10^6 B_z}. \tag{36}$$

In units of gauss, centimeters, volts and amperes, this becomes

$$a_{\text{cm}} = \frac{\sqrt{\xi I\times 10^6}}{1.203V_b^{1/4}B_{z\,(\text{gauss})}} = \frac{\sqrt{\bar{K}V_b}}{1.203B_{z\,(\text{gauss})}}. \tag{36a}$$

All linear dimensions will be expressed in units of a, in accordance with the relations as

$$R_b = \frac{b}{a},\quad R_g = \frac{g}{a},\quad R = \frac{r}{a},\quad \text{and}\quad Z = \frac{z}{a}. \tag{37}$$

We can also normalize the velocity components by defining

$$\gamma_z = \frac{\dot z}{\sqrt{\dfrac{2eV_b}{m}}}\quad\text{and}\quad \gamma_r = \frac{\dot r}{\sqrt{\dfrac{2eV_b}{m}}}. \tag{38}$$

The generalized potential function can now be written

$$\frac{1}{V_b}\left(V - \frac{1}{2}\frac{m}{e}r^2\dot\theta^2\right) = 1 - 0.01515\,\bar{K}\left[\log_e\frac{R_b^2}{R^2} + R^2\left(1 - \frac{R_g^2}{R^2}\right)^2\right] \tag{39}$$

or, on substituting in the Hamiltonian,

$$\gamma_r^2 + \gamma_z^2 = 1 - 0.01515\bar{K}\left[\log_e\frac{R_b^2}{R^2} + R^2\left(1 - \frac{R_g^2}{R^2}\right)^2\right]. \tag{40}$$

Finally the trajectory equation becomes

$$a^2\ddot{R} = \frac{eV_b}{m}\frac{\partial}{\partial R}\frac{1}{V_b}\left[V - \frac{1}{2}\frac{m}{e}r^2\dot{\theta}^2\right]$$

or

$$a^2\ddot{R} = \left(\frac{2eV_b}{m}\right)0.01515\overline{K}\left[\frac{1}{R} - \left(1 - \frac{R_g^4}{R^4}\right)R\right]. \quad (41)$$

From (41) we observe that there is one particular equilibrium radius at which the radial acceleration is zero. Were a beam to be started at this radius with zero radial velocity, the beam could be maintained indefinitely with the same cross section. Designating the equilibrium radius as r_e or in normalized units as R_e where

$$R_e = \frac{r_e}{a},$$

we can solve for R_e by equating the bracketed term in (41) to zero, obtaining

$$R_e = \left[\tfrac{1}{2} + \tfrac{1}{2}\sqrt{1 + 4R_g^4}\right]^{1/2} \quad (42)$$

or if we solve for R_g

$$R_g = (R_e)^{1/2}(R_e^2 - 1)^{1/4}. \quad (42a)$$

When there is no magnetic field at the cathode and hence R_g is equal to zero (we remember that R_g is the radius in normalized units of the flux line threading through the edge of the cathode), the value of R_e is unity and the equilibrium radius reduces to the normalizing constant a. The value of a is then the equilibrium radius for the case in which $R_g = 0$. From (42) we note that R_e varies but slowly with R_g, and from (36) we can see that the actual equilibrium radius varies directly with the square root of the charge density and inversely with the magnetic field.

The potential function in (40) should have a maximum at $R = R_e$ as its derivative is zero at this value. Now let us define a potential function $P(R)$ in the following way which involves the variable R in such a manner that

$$P(R) = -\log_e R^2 + R^2\left(1 - \frac{R_g^2}{R^2}\right)^2$$

$$= -\log_e R^2 + R^2\left[1 - \frac{R_e\sqrt{R_e^2 - 1}}{R^2}\right]^2 \quad (43)$$

$$P(R) - P(R_e)$$

$$= -\log_e\frac{R^2}{R_e^2} + (R^2 - R_e^2)\left[1 - \frac{R_e^2 - 1}{R^2}\right]. \quad (44)$$

$P(R) - P(R_e)$ is a positive function and is zero at $R = R_e$. We can have a power series expansion of this function around $R = R_e$. Writing $R/R_e = X$, we can expand it in terms of a power series of $(1 - X)$ or $(1 - X^2)$, such as

$$P(R) - P(R_e) = 2(2R_e^2 - 1)(1 - X)^2 + \sum_{n=3}^{\infty}\frac{(1 - X)^n}{n}\left[n(n + 1)R_e^2 - n(n + 1) + 2\right]$$

$$= \frac{(R_e^2 - 1)(1 - X^2)}{X^2} + \sum_{n=2}^{\infty}\frac{(1 - X^2)^n}{n}. \quad (45)$$

In terms of the P function we can rewrite (39)

$$\frac{1}{V_b}\left(V - \frac{1}{2}\frac{m}{e}r^2\dot{\theta}^2\right) = \gamma_z^2 + \gamma_r^2$$

$$= 1 - 0.01515\overline{K}\left[\log_e\frac{R_b^2}{R_e^2} + R_e^2\left(1 - \frac{\sqrt{R_e^2 - 1}}{R_e}\right)^2\right]$$

$$- 0.01515\overline{K}[P(R) - P(R_e)]. \quad (46)$$

$P(R) - P(R_e)$ functions are plotted in Fig. 4 with R_e as parameters. A potential trough like this always gives rise to an oscillatory orbit, i.e., the radius will vary from a certain maximum value to a certain minimum value. The magnitude of the maximum or minimum radius depends upon the total energy in the radial direction. The oscillation is analogous to the simple harmonic motion of a simple pendulum in which the amplitude of oscillation depends upon the total energy, except for the fact that our potential trough is unsymmetrical about $R = R_e$ and the amplitude and period are nonlinearly related. The kinetic energy as expressed by γ_r is maximum when the potential energy is minimum, and γ_r is obviously zero when R is at its maximum or minimum values. From (46), the maximum value of γ_r is denoted as γ_{rmax}, which must be equal to the value at which $R = R_e$. We have, therefore,

$$\gamma_{rmax}^2 = 1 - 0.01515\overline{K}\left[\log_e\frac{R_b^2}{R_e^2} + R_e^2\left(1 - \frac{\sqrt{R_e^2 - 1}}{R_e}\right)^2\right] - \gamma_z^2. \quad (47)$$

Since R_b is always greater than R_e if the beam is to pass through the drift tube at all, the quantity in the bracket is always positive. Theoretically then, the maximum kinetic energy depends upon how great a fraction of energy in the z direction, γ_z^2, is retained. The larger γ_z is, the less γ_{rmax} is, and the less the amplitude of oscillation. In the limit, when γ_z is made equal to the quantity preceding it in (47), γ_{rmax}^2 can be made equal to zero. This particular case with $\gamma_{rmax} = 0$ corresponds to the straight rectilinear beam with a radius equal to the equilibrium radius. Thus γ_{rmax} gives a measure of the total energy in the radial direction and can serve as a characteristic constant in specifying a definite orbit. Equally as well, we can use the maximum radius R_{max} or the minimum radius R_{min} to specify a certain

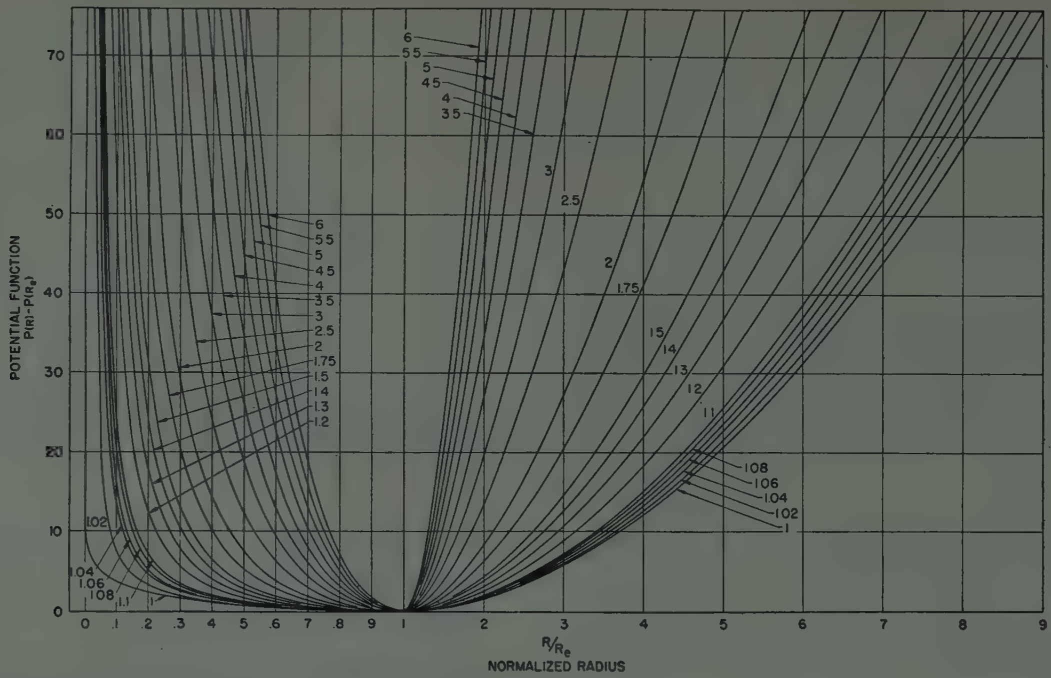


Fig. 4—Normalized potential function

$$P(R) - P(R_e) = -\log_e \frac{R^2}{R_e^2} + (R^2 - R_e^2)\left[1 - \frac{R_e^2 - 1}{R^2}\right]$$

as plotted as a function of R/R_e with R_e as parameter. Different scales of R/R_e are used for $R/R_e<1$ and $R/R_e>1$ to give more accuracy of plotting in the former case.

orbit. They are related to each other by setting $\gamma_r=0$, $R=R_{max}$ or $R=R_{min}$ in (46) substituting (47) in (46). We then have

$$\gamma_{rmax}^2 = 0.01515\overline{K}\left[P\begin{pmatrix}R_{max}\\R_{min}\end{pmatrix} - P(R_e)\right]. \tag{48}$$

As in the simple pendulum problem, in specifying the total energy γ_{rmax}, the orbit is completely determined except for the phase. The particular phase the wave is to pick depends upon the initial conditions. What we can do is to make numerical integrations in (46) for different values of γ_{rmax} and to determine for each case the corresponding values of R/R_e to a normalized axial distance Z measured from a certain point, Z_e. Z_e is defined as the Z co-ordinate where the beam crosses the equilibrium radius i.e., when $R/R_e=1$. In Fig. 5, we see that the beam is having a wave motion of R versus Z. In this figure, we define the successive values of Z_{max}, Z_{min}, and Z_e to correspond with the radius R_{max}, R_{min}, and R_e. It is necessary only to plot the wave in a half-wave section from R_{min} to R_{max} because the wave is a successive inversion around Z_{max} and Z_{min}. With this understanding, the numerical integration is given as a function of R/R_e versus $Z-Z_e$. To do this, we notice,

$$\frac{dR}{dZ} = \frac{\gamma_r}{\gamma_z}, \qquad \left(\frac{dR}{dZ}\right)_{max} = \frac{\gamma_{rmax}}{\gamma_z}, \tag{49}$$

and Z_e is taken as the one closest to it.

Substituting (49) and (47) into (46), we have

$$\left(\frac{dR}{dZ}\right)^2 = \left(\frac{\gamma_r}{\gamma_z}\right)^2 = \left(\frac{dR}{dZ}\right)_{max}^2 - \frac{0.01515\overline{K}}{\gamma_z^2}\left[P(R) - P(R_e)\right] \tag{50}$$

$$\frac{\sqrt{\overline{K}}}{\gamma_z}\,\frac{Z - Z_e}{R_e} = \int_{R_e}^{R} \frac{\pm d\left(\dfrac{R}{R_e}\right)}{\sqrt{0.01515}\left\{\dfrac{\left(\dfrac{dR}{dZ}\right)_{max}^2 \gamma_z^2}{0.01515\overline{K}} - \left[P(R) - P(R_e)\right]\right\}^{1/2}}. \tag{51}$$

Equation (51) is integrated numerically and the results are plotted in Fig. 6. Each plate corresponds to a certain R_e and the parameters are

$$\gamma_z\left(\frac{dR}{dZ}\right)_{max}\frac{1}{\sqrt{0.01515\overline{K}}}.$$

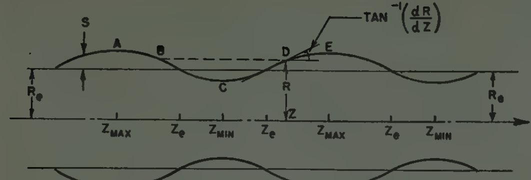
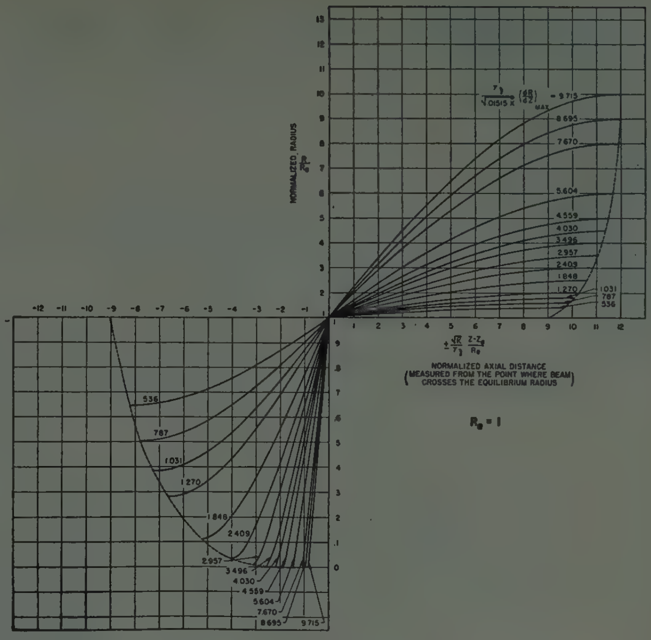


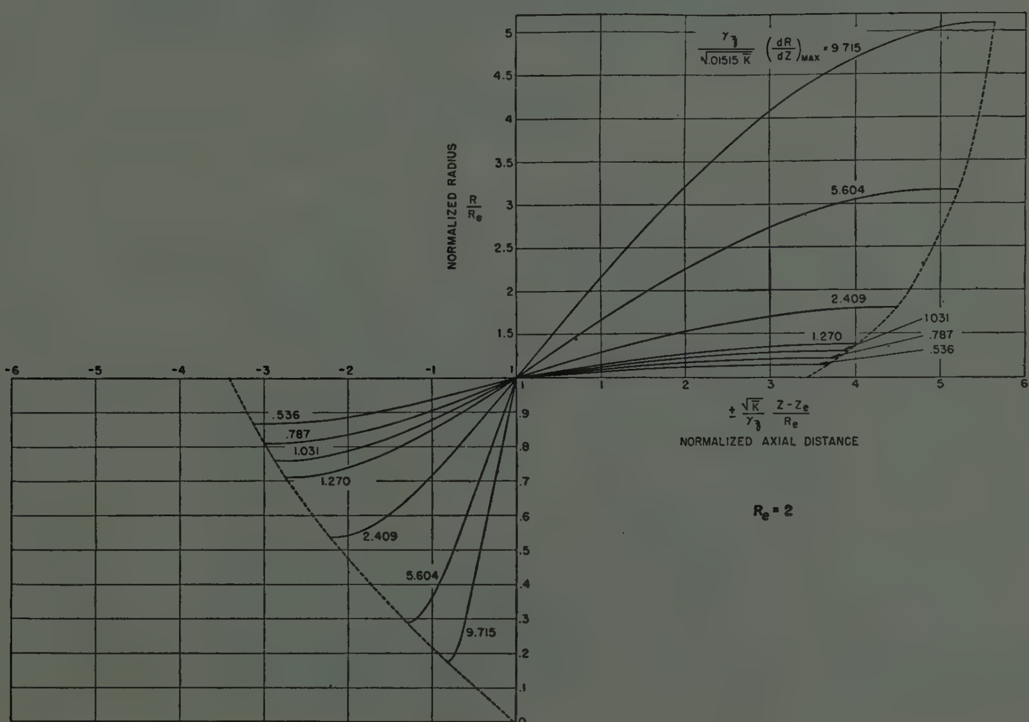
Fig. 5—Diagrams illustrating the definitions of the quantities Z_{max}, Z_{min}, Z_e, dR/dZ, R_{max}, R_{min}, R_e, and S of wave motion of a beam.

In Fig. 7, the corresponding values of dR/dZ are plotted versus $Z-Z_e$ for the same parameters. The sign to be adopted in these diagrams is determined by whether the beam is increasing in amplitude or decreasing in amplitude. Thus at D, in Fig. 5 dR/dZ is positive, while at B it will be exactly the same in magnitude but negative in sign. Similarly, $Z-Z_e$ at B is positive, while that at D is equal and opposite in sign.

The "quarter wavelength" $Z_{max}-Z_e$ and $Z_{min}-Z_e$ are plotted as a function of $(dR/dZ)_{max}$ in Fig. 8 and, similarly, the "half wavelength" $Z_{max}-Z_{min}$ is plotted in Fig. 9. Through use of these curves, the wave can be extended to any number of wavelengths as desired.



(a)



(b)

Fig. 6—Actual numerically integrated curves of the wave motions in a uniform electrostatic and magnetic field with R/R_e plotted against normalized axial co-ordinate

$$\pm\frac{\sqrt{K}}{\gamma_s}\left(\frac{Z-Z_0}{R_e}\right).$$

Each plate is plotted for a specified value of R_e. The parameters

$$\frac{\gamma_s}{\sqrt{0.01515K}}\left(\frac{dR}{dZ}\right)_{max}$$

are the normalized maximum slope of the wave motion and are the characteristics of different possible orbits. Notice that a nonuniform scale of R/R_e is used for $R/R_e>1$ and $R/R_e<1$ to give more details for the latter cases. The dotted lines are the locus for maximum and minimum points of the wave.

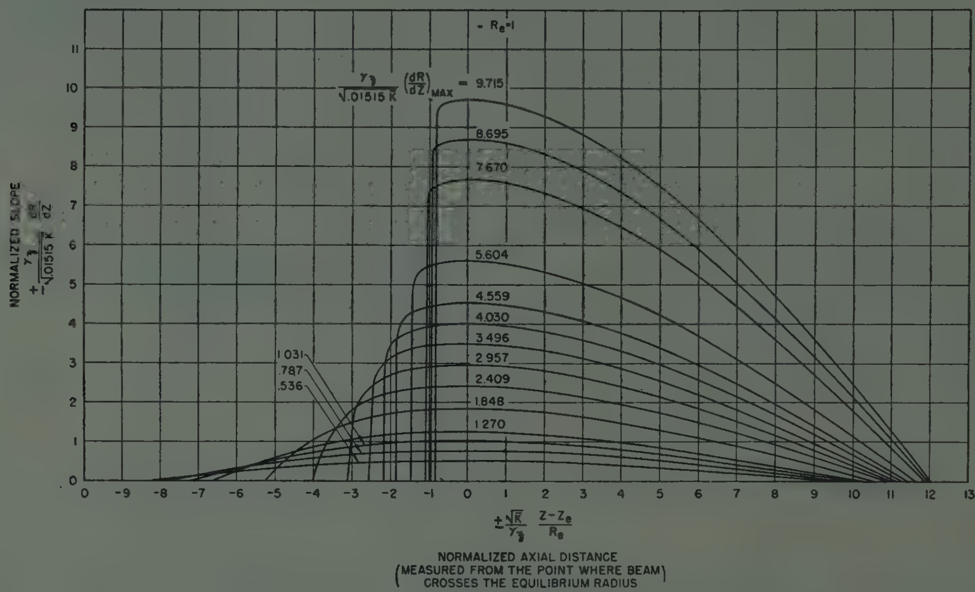
To explain these curves, we rewrite (41) by defining another variable S as the deviation of R from R_e; thus

$$S = R - R_e$$
$$\frac{d^2R}{dZ^2} = \frac{d^2S}{dZ^2}. \qquad (52)$$

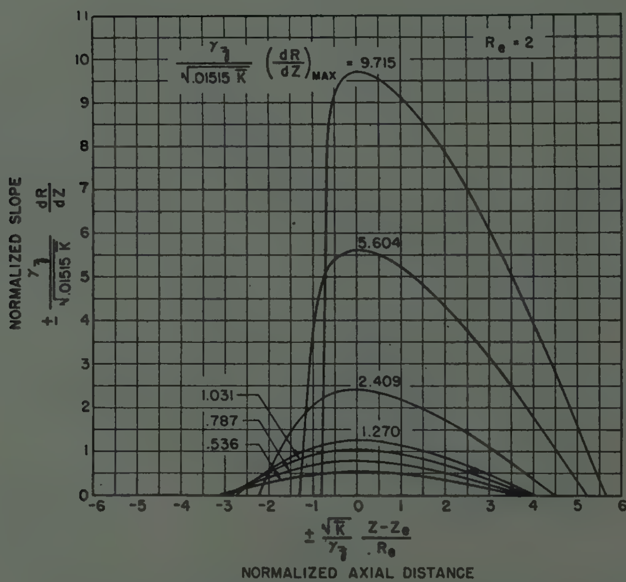
In the function at the right-hand side of (41) we can factor $(R-R_e)$ out and replace it by S, but leave R in the rest of this factor. After substitution from (52), we have

$$\frac{d^2S}{dZ^2} + \frac{0.01515\overline{K}}{\gamma_z^2}\left(1+\frac{R_e}{R}\right)\left(1+\frac{R_e^2-1}{R^2}\right)S = 0. \qquad (53)$$

This is a perfect simple wave equation of S, if the coefficient of S is a constant quantity. Indeed, we have a simple sine wave about $R=R_e$ when S remains small and



(a)



(b)

Fig. 7—The corresponding value of the slope of the wave motion

$$\pm\frac{\gamma_z}{\sqrt{0.01515\overline{K}}}\frac{dR}{dZ}$$

is plotted versus

$$\pm\frac{\sqrt{\overline{K}}}{\gamma_z}\frac{Z-Z_e}{R_e}$$

with the same parameters and R_e values as in Fig. 6.

R can be set equal to R_e in the coefficient of S in (53). Thus, for small S, the normalized wavelength λ/a must be

$$\lim_{S\to 0} \frac{\lambda}{a}\,\frac{\sqrt{K}}{\gamma_z} = \frac{\pi}{\sqrt{0.01515}\left(1-\frac{1}{2R_e^2}\right)^{1/2}}. \tag{54}$$

The solution for small perturbation is then simply

$$S = S_{max}\sin\left[\frac{2\pi a}{\lambda}Z+\alpha\right] \tag{55}$$

where S_{max} and α depend upon the initial conditions. We see from Fig. 6 that the wave is nearly symmetrical and sinusoidal for small values of $(dR/dZ)_{max}$. Equation (54) actually gives the asymptotic values of wavelentghs for $(dR/dZ_{max}=0$, as appeared in Figs. 6 to 8. Now as S gets large, we might still consider (53) as the wave equation, but with a variable wavelength as the radius R varies. Since R appears in the denominators of the coefficient in (53), the "local wavelength" increases as the radius increases. Thus, in Fig. 8, the quarter wavelength $Z_{max}-Z_e$ increases as $(dR/dZ_{max}$ increases, while that of $Z_{min}-Z_e$ decreases as $(dR/dZ_{max}$ increases. In the limit, when $(dR/dZ)_{max}$ is very large, the quarter wavelength will approach the following asymptotic values as R approaches ∞ and 0, respectively,

$$\lim_{R/R_e>>1}\frac{\sqrt{K}}{\gamma_z}\left|Z_{max}-Z_e\right| = \frac{\pi}{2\sqrt{0.01515}} \tag{56}$$

$$\lim_{R/R_e>>1}\frac{\sqrt{K}}{\gamma_z}\left|Z_{min}-Z_e\right| = 0. \tag{56a}$$

Therefore, for large perturbation, the wavelength will approach twice the value corresponding to (56), or

$$\lim_{(dR/dZ)_{max}\ \text{large}}\frac{\lambda}{a}\,\frac{\sqrt{K}}{\gamma_z} = \frac{\pi}{\sqrt{0.01515}}. \tag{57}$$

We see that for $R_e=1$, the wavelength for a small perturbation is $\sqrt{2}$ times as large as that for the large perturbation. For large R_e's, the difference becomes very slight.

Using equation (36), we can eliminate a and $\sqrt{K}$ in (56) and (57), showing that the wavelength is actually directly related to the magnetic field used. Thus,

$$\lim_{S\to 0}\left(\frac{\lambda f}{\dot z}\right)\left(\frac{c}{f}\right)_{cm} B_{gauss} = 10{,}700\,\frac{1}{\sqrt{1-\frac{1}{2R_e^2}}} \tag{58}$$

$$\lim_{(dR/dZ)_{max}\ \text{large}}\left(\frac{\lambda f}{\dot z}\right)\left(\frac{c}{f}\right)_{cm} B_{gauss} = 10{,}700, \tag{59}$$

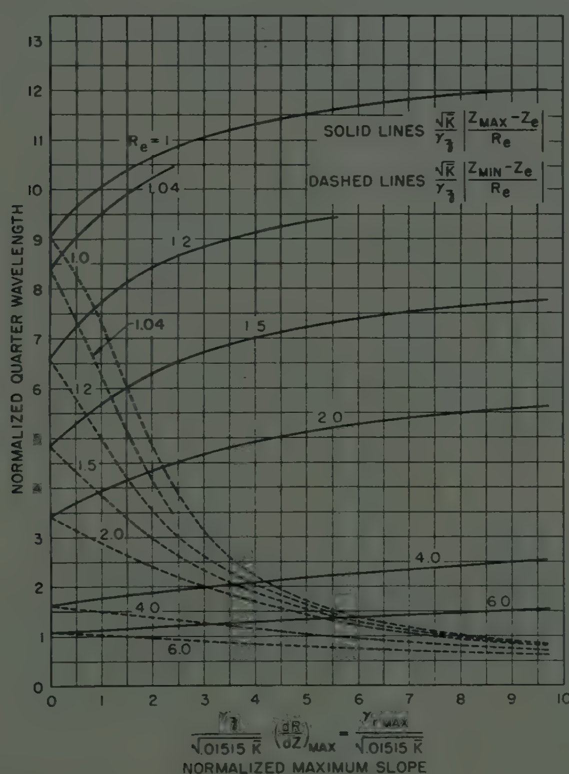


Fig. 8—Values of "quarter wavelength"

$$\frac{\sqrt{K}}{\gamma_z}\left|\frac{Z_{max}-Z_e}{R_e}\right| \quad \text{and} \quad \frac{\sqrt{K}}{\gamma_z}\left|\frac{Z_{min}-Z_e}{R_e}\right|$$

are plotted as a function of the characteristic constant

$$\frac{\gamma_z}{\sqrt{0.01515K}}\left(\frac{dR}{dZ}\right)_{max}$$

for different values of R_e.

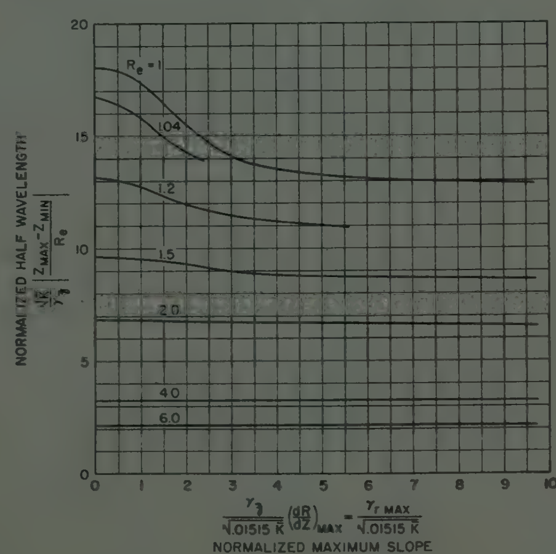


Fig. 9—Values of "half wavelength"

$$\frac{\sqrt{K}}{\gamma_z}\left|\frac{Z_{max}-Z_{min}}{R_e}\right|$$

are plotted as a function of the characteristic constant

$$\frac{\gamma_z}{\sqrt{0.01515K}}\left(\frac{dR}{dZ}\right)_{max}$$

for different values of R_e.

where f is arbitrarily inserted because it is actually cancelled out between the two factors. The first factor measures the wavelength in terms of number of cycles of transit of the frequency under consideration. c/f is the wavelength in cm.

Variable Fields

According to equations (25) and (26), the best way to handle the variable field case—and the exact way—is to construct a mechanical potential model of $V-\frac{1}{2}(m/e)r^2\dot{\theta}^2$ and run different kinds of trajectories with steel balls. A complete model can be constructed even involving changes of electrostatic potentials such as those in the pre-accelerating region. However, the general analytic solution with both variable electric and magnetic fields are very much involved; we choose to treat the constant Laplacian electric field again but to have only the magnetic field variable. There are many applications in which the magnetic field is not constant. Such is the case, for example, when the magnetizing coil is not a perfect solenoid and has to be localized to yield spaces for electromagnetic field structures. Also, in order to introduce rotation into the beam, the beam has to cross the radial magnetic field somewhere in an intermediary region. That is the place where the magnetic field is necessarily varying.

To construct a variable magnetic field potential model, we can start from (46). Although this is written for a constant magnetic field case, it can be extended to the variable magnetic field case by considering a as a variable, according to (36), in which B_z can be considered as the magnetic field along the axis and is a function of z. We see that a is a quantity that varies inversely as the magnetic field B_z. Similarly, g also should be inversely proportional to the square root of the magnetic field. From (42) we can calculate R_e, the normalized equilibrium radius, as a function of distance. The potential function $P(R)-P(R_e)$ can be constructed at different points along the z axis. The potential surface is formed by connecting the potential troughs smoothly together. Thus, in a constant field the model would be a cylindrical surface formed by one of the potential troughs of Fig. 4, while, with variable fields, the surface is a distorted form of cylindrical surfaces consisting of successive troughs of Fig. 4 with variable equilibrium radius. In particular, when there is no magnetic field, the equilibrium radius is infinite, and the potential curve decreases logarithmically as R increases. An actual potential model made of plaster of Paris and paper is shown in Fig. 10, in which a logarithm potential region C for pure electrostatic field is connected to a cylindrical trough A for a constant magnetic field. The intermediate region B is the variable magnetic field region.

Analytically, for very slow variation of magnetic field, we can take the first-order expansion of the magnetic potential. This is the same if we treat a (which is understood to vary inversely as B_z at the axis) as varying with z. Similarly, g varies as $\sqrt{a}$. For completeness,

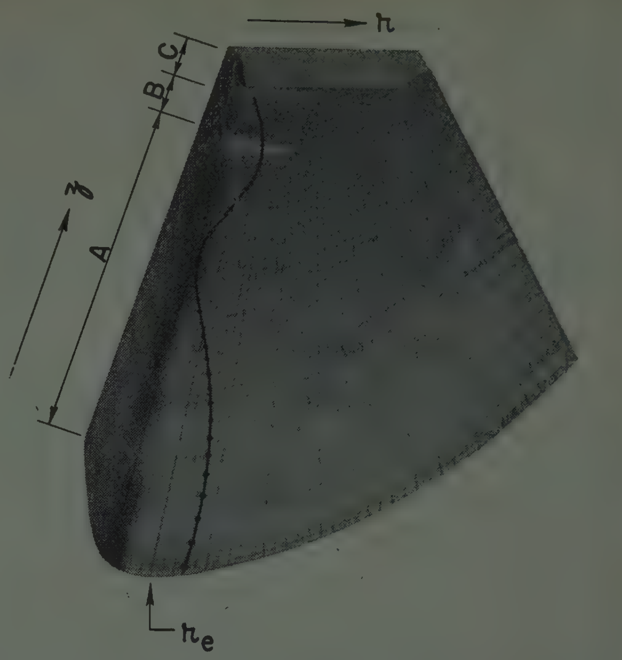


Fig. 10—Mechanical equivalent potential model illustrating the potential functions in the r-z plane with different values of magnetic field and constant Laplacian electric field. The cylindrical trough, section A, represents a section with uniform magnetic field. The logarithmic potential, section C, represents the pure electrostatic case without magnetic field. The skewed surface B connecting section A with section C represents the change of potential function where the magnetic field is injected into the beam and the beam starts rotation. Round dots are steel balls which are made to run in the model, and simulate the paths that would be traced by a typical electron beam.

we can also consider b, the radius of the drift tube, as varying with z. Then we have

$$\begin{aligned} \frac{m}{2eV_b}\ddot{r} &= 0.01515\overline{K}\left[\frac{1}{r}-\frac{r}{a^2}\left(1-\frac{g^4}{r^4}\right)\right] \\ \frac{m}{2eV_b}\ddot{z} &= 0.01515\overline{K}\left[\frac{1}{b}\frac{db}{dz}-\frac{r^2}{a^2}\left(1-\frac{g^2}{r^2}\right)\frac{1}{a}\frac{da}{dz}\right]. \end{aligned} \tag{60}$$

Note that the trajectory differential equation relating r to z takes the form

$$\dot{z}^2\frac{d^2r}{dz^2} = \ddot{r} - \ddot{z}\frac{dr}{dz}. \tag{61}$$

We see that (60) can be substituted into (61) as the coefficient of dr/dz and also the fixed term. If $\dot{z}$ can be considered constant, or nearly so, it is a second order differential equation. Neglecting the term involving dr/dz, we see that the curvature is negative when r is greater than the equilibrium radius and vice versa, so that the wave has a tendency of always bending toward the equilibrium radius. Equation (60) shows that as a result of the variation of the magnetic field or the drift tube size, the beam is accelerated or retarded in the z direction as well. It is retarding when the magnetic field is increasing or the drift tube wall is enlarging, and vice versa. A negative acceleration $\ddot{z}$ and a positive slope dr/dz would add a more positive curvature, and there are different possible combinations that can actually affect the trajectory. In general, the wave still has a tendency to curve around the equilibrium radius.

The correction term in (61) tends to increase the angle of inclination when $\ddot{z}$ is negative or retarding. This is understandable when energy is converted from the z direction into other forms. As in an increasing magnetic field, the $\dot{z}$ component is decreasing, and a larger angle results if everything else remains the same.

Another extreme case is obtained when the magnetic field does not vary slowly, but rather has a sudden transition. Such is the case, for example, when the field is applied to the beam through a magnetic pole piece. The discontinuity of horizontal component B_z is brought about by an infinite B_r at the boundary, for the magnetic flux must enter the beam in an infinitesimal distance. Under this condition, this infinite B_r reacts with $\dot{z}$ and through Lorentz force, taking energy in or out of the rotational component $r\dot{\theta}$. Conversely, B_r reacts with $r\dot{\theta}$ to add or subtract energy from $\dot{z}$. In this manner there is perfect interchange of energy between $\dot{z}$ and $r\dot{\theta}$. On the other hand, $\dot{r}$ and r cannot change appreciably in reacting with a finite magnetic field. B_z in a short interval. Thus, at the boundary, we must have

$$\begin{aligned} \dot{r}\big|_+ &= \dot{r}\big|_-, \qquad \gamma_z \frac{dr}{dz}\bigg|_+ = \gamma_z \frac{dr}{dz}\bigg|_- \\ r\big|_+ &= r\big|_-, \end{aligned} \tag{62}$$

Now if on both sides of the boundary the field is constant, we can find out the value of a at the left or right of the boundary. By using continuation equations of (62) we can carry the initial conditions to the second region from which the trajectory can follow that in the previous section. It takes a few trials because γ_z is different on both sides and depends upon dr/dz and r/a in each case, but this procedure is not difficult, especially when γ_z is close to unity.

This step-function method can be used to analyze a slowly varying field case as well by approximating a continuously varying magnetic field with a step varying field. The process described automatically takes care of the variation of the $\ddot{z}$ terms in (61) if we make fine enough divisions in the step functions.

Relativity Correction

When we operate a beam at very high voltage, the relativity corrections must be applied. In general, the self-magnetic field tends to compensate for the electrostatic space charge repulsion force, so that for the same equilibrium radius the magnetic field required is less than proportional to the square root of voltage as indicated in (36). Let us define a quantity

$$\sigma = 1 + \frac{eV}{mc^2}\cdot$$

It can be shown that the normalization constant a of (36) must be modified so that

$$a_{cm} = \frac{\sqrt{KV_b}}{1.203 B_{z\ \text{gauss}}}\left(\frac{2}{\sigma+1}\right)^{1/4}. \tag{63}$$

If we again normalize everything with this new value of a and make a new transformation,

$$Z' = \left(\frac{2}{\sigma+1}\right)^{3/4} Z. \tag{64}$$

Equation (51) again takes the form

$$\frac{\sqrt{K}}{\gamma_z}\frac{Z'-Z_e'}{R_e} = \int_{Re}^{R} \frac{\pm\left(\dfrac{R}{R_e}\right)}{\sqrt{0.01515}\left[\dfrac{\left(\dfrac{dR}{dZ'}\right)^2_{\max}\gamma_z^2}{0.01515\bar{K}} - [P(R)-P(R_e)]\right]^{1/2}} \tag{65}$$

where γ_z is defined as

$$\gamma_z = \frac{\dot{z}}{\sqrt{\dfrac{2eV_b}{m}}}\sqrt{\frac{\sigma+1}{2\sigma^2}}\cdot \tag{66}$$

The curves of Figs. 6 through 9, can again be used with the understanding that Z is replaced by Z'.

The net result can thus be summarized as follows:

(1) The magnetic field required to give a parallel beam is less with relativity correction by a factor $(2/(\sigma+1))^{1/4}$ compared with the case without correction.

(2) In the case where the beam is not strictly parallel, the wave motion is such that comparing the same perveance and the same ratio of maximum radius to the equilibrium radius, the wavelength appears to be longer by a factor of $((\sigma+1)/2)^{3/4}$.

It should also be noted that the perveance at high voltage with space-charge limited operation is not the same as that at the lower voltage for the same tube. It is generally lower. Based upon a plane diode, the ratio of actual perveance K to that at low voltage K_0 has the following relations

$$\begin{aligned} \frac{K}{K_0} &= \frac{9}{8\sqrt{2}}\,\frac{\displaystyle\int_1^\sigma \frac{dy}{(y^2-1)^{1/4}}}{(\sigma-1)^{3/2}} \\ &= \left[1+\frac{\sigma-1}{2}\right]^{3/2}\Bigg[1-\frac{3}{14}(\sigma^2-1) \\ &\quad + \frac{9}{88}(\sigma^2-1)^2 - \frac{1}{16}(\sigma^2-1)^3 \\ &\quad + \frac{105}{2432}(\sigma^2-1)^4 \cdots\Bigg]^2 \\ &\doteq 1 - \frac{3}{28}(\sigma-1). \end{aligned} \tag{67}$$

Thus the true perveance as corrected by (67) must be used in (66); this correction will again make the wavelength even longer for the same tube, compared with low voltage operation.

The Transmission and Reception of Elliptically Polarized Waves*

GEORGE SINCLAIR†, SENIOR MEMBER, IRE

Summary—**A vector parameter is defined which represents a generalization of the effective length of an antenna to include a specification of the polarization of the field radiated by the antenna. It is shown that the parameter so defined is also useful in calculating the voltage at the terminals of the antenna when it is used to receive plane waves of arbitrary (elliptical) polarization.**

I. Introduction

ALTHOUGH IT HAS been known for some time that the radiation from many antennas, particularly airborne antennas, is elliptically polarized, comparatively little attention has been given to this aspect of the radiation. When measurements are made to determine the patterns of such antennas, the radiation is usually treated as if it were linearly polarized. The electric field intensity vector is resolved into two orthogonal components in a suitable coordinate system, but the fact that these two components may not be in time phase is usually ignored.

In recent years there has been increasing interest in antennas designed to radiate circularly polarized waves and it has become necessary to devise methods for specifying the performance of such antennas. The parameters commonly employed to characterize the properties of a transmitting antenna (gain, effective length, equivalent aperture, etc.) require modification when applied to systems using circular or elliptical polarization. The need for modification of the parameters becomes particularly apparent in determining the performance of such antennas when receiving waves of arbitrary polarization characteristics.

It is desirable, therefore, to examine carefully the properties of systems which radiate elliptically polarized waves, and to devise convenient methods for measuring and specifying the performance of these systems. There is a need for the definition of a parameter to specify the performance of the system, whether transmitting or receiving, and which includes a specification of the polarization.

II. The Transmission of Elliptically Polarized Waves

The problem of defining a parameter to characterize the performance of antennas which radiate elliptically polarized waves has been considered by Burgess,[1] who has defined a *generalized effective length* $h(\theta, \phi, \alpha)$ by means of the equation[2]

$$E(\alpha) = \frac{Z_0 I h}{2\lambda r}, \tag{1}$$

where

$E(\alpha)$ = the electric field component of polarization α
r ≐ distance from the antenna to the point where E is measured
I = the terminal current
Z_0 = the intrinsic impedance of space for plane waves ($120\,\pi$ ohms)
λ = wavelength.

The angles θ and ϕ are the usual co-ordinate angles in a spherical system (see Fig. 1).

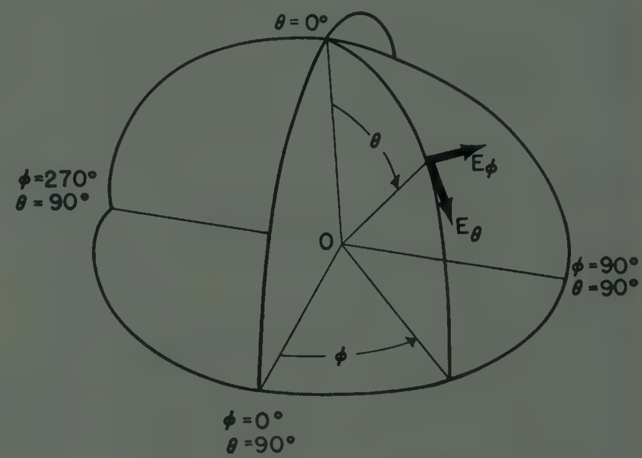


Fig. 1—Resolution of the field into components in a spherical co-ordinate system.

This method of defining the effective length of an antenna yields a parameter which is useful, not only for specifying the performance of a transmitting antenna, but also for specifying the performance when the antenna is receiving plane waves of arbitrary linear polarization. However, it is not too useful in the situation when the antenna is receiving plane waves of arbitrary elliptical polarization.

Another parameter, which has been used to specify the performance of transmitting antennas of arbitrary polarization characteristics, is the radiation vector $\boldsymbol{N}$ used by Schelkunoff.[3] The radiation vector is equal to

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[1] R. E. Burgess, "Aerial characteristics," *Wireless Eng.*, vol. 21, pp. 145–160; April, 1944.
[2] Rationalized mks units are used throughout.
[3] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., New York, N. Y., pp. 332–335; 1943.

the principal term in the expression for the magnetic vector potential $\boldsymbol{A}$ at large distances from the antenna. The electric field intensity $\boldsymbol{E}$ is related to $\boldsymbol{N}$ by the equation

$$\boldsymbol{E} = -j\frac{Z_0\boldsymbol{N}}{2\lambda r}e^{-jkr}. \tag{2}$$

The factor $e^{j\omega t}$ containing time is implied.

The radiation vector can be a complex vector, so it is capable of characterizing a field of arbitrary elliptical polarization and, therefore, it completely specifies the distant field of an arbitrary transmitting antenna. Since $\boldsymbol{N}$ is proportional to the current amplitude in the antenna when transmitting, it must be modified if it is to be applied to receiving antennas. By dividing the vector $\boldsymbol{N}$ by the amplitude of the input current to the antenna, a parameter is obtained which is independent of the amplitude of the current, and which depends only on the physical configuration of the antennas and on the method of feeding. Let

$$\boldsymbol{h} = \frac{-\boldsymbol{N}}{I}, \tag{3}$$

so that (2) becomes

$$\boldsymbol{E} = j\frac{Z_0I\boldsymbol{h}}{2\lambda r}e^{-jkr}. \tag{4}$$

In (3) the negative sign has been inserted so that in the case of linear antennas the vector $\boldsymbol{h}$ will reduce to the usual effective length.

The vector $\boldsymbol{h}$ is obviously a complex vector and it has the dimensions of length. It is only in the special case of linear antennas that a convenient physical interpretation can be given to the fact that it has the dimensions of length. In some respects it would be better to define the vector in a way which would make it dimensionless. This could be done, for example, by combining the wavelength λ (which appears in the denominator of (4)) with $\boldsymbol{h}$, thus obtaining a dimensionless parameter

$$\boldsymbol{h}' = \frac{\boldsymbol{h}}{\lambda}, \tag{3a}$$

so that

$$\boldsymbol{E} = j\frac{Z_0I\boldsymbol{h}'}{2r}e^{-jkr}. \tag{4a}$$

However, in view of the widespread use of the concept of effective length, the definition in (3) will be retained in the following. The vector $\boldsymbol{h}$ is a function of the co-ordinate angles θ and ϕ, but does not depend on the radial co-ordinate r (because, by definition, $\boldsymbol{N}$ does not depend on r). Equation (4) gives only the distant field from the antenna, so that $\boldsymbol{E}$ and $\boldsymbol{h}$ have only θ components and ϕ components.

For a short linear element of length $2L$, oriented so its axis coincides with the Z axis (Fig. 1) and carrying a uniform current, it is readily found that $\boldsymbol{h}$ has only a θ component

$$h_\theta = L\sin\theta. \tag{5}$$

For a half-wave dipole antenna, parallel to the Z axis,

$$h_\theta = \frac{\lambda}{\pi}\,\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}. \tag{6}$$

In the case of a small current loop of area S, with the axis of the loop parallel to the Z axis, it is found that $\boldsymbol{h}$ has only a ϕ component

$$h_\phi = -j\frac{2\pi S}{\lambda}\sin\theta. \tag{7}$$

Thus $\boldsymbol{h}$ is a negative imaginary quantity and of magnitude equal to the effective length of a loop as ordinarily defined.

In general, $\boldsymbol{h}$ may have both θ and ϕ components and both may be complex. However, one of these components (say the θ component) can be made a positive real quantity by a suitable choice of the origin for time, and it will usually be convenient to do this. The ϕ component of $\boldsymbol{h}$ can then be written in the form

$$h_\phi = |h_\phi|\,e^{j\delta_1}, \tag{8}$$

where δ_1 is the phase angle by which the ϕ component of the field leads the θ component. The phase angle δ_1 may lie in any one of the four quadrants.

For an antenna which radiates circularly polarized waves, the vector $\boldsymbol{h}$ can be resolved into two components equal in magnitude with one component lagging or leading the other by 90° in time phase. That is

$$h_\phi = h_\theta e^{j\delta_1}, \tag{9}$$

where δ_1 is $\pm 90°$. If $\delta_1 = +90°$, then the electric vector appears to rotate in the clockwise direction (see Fig. 2). If $\delta_1 = -90°$, the rotation is counterclockwise.[4]

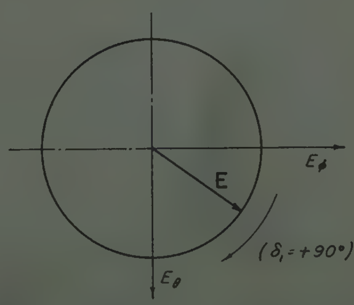

Fig. 2—The transmitted field, as seen by an observer viewing the source of the field.

[4] The rotation specified is that seen by an observer looking toward the source of the wave. These rotations are often specified as right-handed or left-handed, but since there are two different conventions in use, it is felt that the present method of specifying the rotation will be less confusing.

III. The Reception of Elliptically Polarized Waves

In the receiving case, the problem to be considered is that of determining the open-circuit voltage produced at the terminals of an antenna of arbitrary polarization characteristics, when receiving a plane wave of arbitrary polarization.[5–7] Consider an antenna which, when transmitting, is characterized by a vector $\boldsymbol{h}$ as defined above. Suppose that this antenna is used to receive a plane wave incident from a direction defined by the angles θ_0 and ϕ_0, and given by the equation

$$\boldsymbol{E}^i = \boldsymbol{E}_0 e^{j\boldsymbol{k}\cdot\boldsymbol{r}}, \tag{10}$$

where

$\boldsymbol{E}_0$ = a constant complex vector
$\boldsymbol{k}$ = the propagation vector[8]
$\boldsymbol{r}$ = radius vector in any direction defined by the angles θ and ϕ

so that

$$\boldsymbol{k}\cdot\boldsymbol{r} = kr[\sin\theta_0 \sin\theta \cos(\phi - \phi_0) + \cos\theta_0 \cos\theta]. \tag{11}$$

The vector $\boldsymbol{E}^i$ may be resolved into its three spherical components at each point in space. However, along the radius vector which defines the direction of incidence of the plane wave, $\boldsymbol{E}^i$ will have only θ and ϕ components and, if the wave is elliptically polarized, these components will not be in time phase. For this particular direction,

$$\boldsymbol{E}^i = (E_{0\theta}\boldsymbol{i}_\theta + E_{0\phi}\boldsymbol{i}_\phi)e^{jkr_1}, \tag{12}$$

where $\boldsymbol{i}_\theta$ and $\boldsymbol{i}_\phi$ are unit vectors and r_1 is distance measured along the direction of propagation. The positive exponent is used in (12) because the wave is an incoming wave. By a suitable choice of origin for time in (12) it is possible to make E_{0_θ} real and positive. Then $E_{0\phi}$ may be complex,

$$E_{0\phi} = |E_{0\phi}| e^{-j\delta_2}, \tag{13}$$

where δ_2, is the time phase angle by which the ϕ component lags the θ component.

Equation (13) should be compared with (8) and it should be noted that there is a difference in sign in the exponential terms. The signs in these equations are arbitrary, and were chosen so that when δ_1 and δ_2 lie in the same quadrant, the two equations (8) and (13) represent the *same* direction of rotation of the vector for an observer viewing the source of the wave in each case (see Figs. 2 and 3). The difference in sign required to do this arises from the difference in signs of the exponentials in (4) and (12) representing an outgoing and incoming wave, respectively.

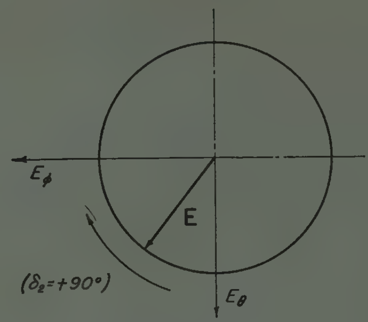

Fig. 3—The incoming plane wave, as seen by an observer viewing the source of the wave.

In determining the open-circuit terminal voltage produced by the incident wave, the θ and ϕ components in the system can be considered separately and the results added to obtain the total voltage. The incident wave expressed by (12) can be seen to consist of two linearly polarized plane waves with a time phase δ_2 between them.

The plane wave component E_{0_θ} will produce a terminal voltage

$$V' = h_\theta E_{0_\theta}. \tag{14}$$

This is readily proved[9] by applying the Reciprocity Theorem, it being supposed that the field E_{0_θ} is produced by a linear dipole of known characteristics, and oriented parallel to E_{0_θ}. Similarly, for the ϕ components only,

$$V'' = h_\phi E_{0\phi} = |h_\phi| |E_{0_\theta}| e^{j(\delta_1 - \delta_2)}. \tag{15}$$

In (15) both h_ϕ and $E_{0\phi}$ are complex numbers (see (8) and (13)). The total terminal voltage is

$$V = V' + V'' = h_\theta E_{0_\theta} + h_\phi E_{0\phi}. \tag{16}$$

It is apparent that (16) can be written as the scalar product of the vectors $\boldsymbol{h}$ and $\boldsymbol{E}_0$,

$$V = \boldsymbol{h}\cdot\boldsymbol{E}_0. \tag{17}$$

Equation (17) represents a generalization of the scalar equation usually written for linearly polarized systems.[10]

IV. Some Applications

For numerical calculations, the following form of (17) is more useful:

$$V = h_\theta E_{0_\theta} + |h_\phi| |E_{0\phi}| e^{j(\delta_1 - \delta_2)}, \tag{18}$$

[5] Yung-Ching Yeh, "The received power of a receiving antenna and the criteria for its design," Proc. I.R.E., vol. 37; pp. 155–158; February, 1948.
[6] W. Sichak and S. Milazzo, "Antennas for circular polarization," Proc. I.R.E., vol. 36, pp. 997–1001; August, 1948.
[7] C. W. Harrison, Jr., and Ronold King, "The receiving antenna in a plane-polarized field of arbitrary orientation," Proc. I.R.E., vol. 32, pp. 35–48; January, 1944.
[8] J. A. Stratton, "Electromagnetic Waves," D. Van Nostrand Co., New York, N. Y., p. 408; 1941.
[9] See page 479 of footnote reference 3.
[10] F. E. Terman, "Radio Engineers Handbook," McGraw-Hill Book Co., New York, N. Y., p. 785; 1943.

where h_0 and E_{0_θ} are assumed to be positive real quantities. A number of well known results can be obtained from this equation. Consider an antenna which, if transmitting, would radiate circularly polarized waves in a given direction; and used to receive circularly polarized waves from the same direction. Then

$$h_\theta = | h_\phi | = h, \tag{19}$$

$$E_{0_\theta} = | E_{0\phi} | = E_0, \tag{20}$$

and

$$V = hE_0(1 + e^{j(\delta_1 - \delta_2)}). \tag{21}$$

If the antenna and the incident wave are characterized by the same direction of rotation, then $\delta_1 = \delta_2 = \pm 90°$,

$$V = 2hE. \tag{22}$$

On the other hand, if the rotations are in opposite directions, $\delta_1 = -\delta_2 = \pm 90°$ so that

$$V = 0. \tag{23}$$

When investigating the nature of the polarization of the field radiated in a given direction by an antenna, measurements are often made to determine the *polarization characteristic* by receiving the field with a linear dipole which is rotated in a plane parallel to the wavefront. The voltage at the terminals of the dipole, when plotted against the *polarization angle* α (the angle between the axis of the dipole and a reference direction) yields the polarization characteristic. It is often assumed that this curve should be an ellipse. The equation of the curve is readily obtained from an application of (17) and by making use of the Reciprocity Theorem. Assume that the dipole is used to transmit, producing a field, near the antenna, which is essentially a plane wave,

$$\mathbf{E}_0 = (E_0 \cos \alpha \mathbf{i}_\theta + E_0 \sin \alpha \mathbf{i}_\phi), \tag{24}$$

where α = angle between the axis of the dipole and the direction of $\mathbf{i}_\theta$.

The received voltage is

$$\begin{aligned} V &= \mathbf{h} \cdot \mathbf{E}_0 \\ &= E_0[h_\theta \cos \alpha + | h_\phi | e^{j\delta_1} \sin \alpha]. \end{aligned} \tag{25}$$

The magnitude of this voltage is

$$\begin{aligned} | V | = E_0[(h_\theta \cos \alpha + | h_\phi | \cos \delta_1 \sin \alpha)^2 \\ + (| h_\phi | \sin \delta_1 \sin \alpha)^2]^{1/2}. \end{aligned} \tag{26}$$

Considering $|V|$ as a function of a polar angle α, it is readily shown that this is not the equation of an ellipse.

The Magnetic Amplifier*

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Summary—The "small signal" theory of the magnetic amplifier is developed under certain simplifying assumptions. Expressions for the amplification are derived in terms of electrical and magnetic quantities, and conditions for optimum amplification are obtained. The predictions of the theory are found to agree qualitatively with experimental results of other workers.

I. Introduction

THE MAGNETIC amplifier or, as more commonly known, the dc excited iron-core reactor, has already been described in the literature. Early descriptions are those of Boyajian,[1] Dällenback and Gerecke[2] and more recently those of Buckhold,[3] Uno Lamm,[4] and Kirschbaum and Harder.[5] It is only in the recent publications that the name "magnetic amplifier" has been used, and Buckhold appears to be the first to derive an expression for the amplification in terms of the characteristics of the magnetic material.

In these treatments the development of the theory is based on the normal magnetization curve. In the following discussion, on the other hand, the displaced hysteresis loop is taken as the starting point. In so doing, emphasis is placed on the instantaneous rather than the average behavior of the amplifier.

In this work, it is assumed (a) that the various quantities (i.e., flux, magnetic intensity, current, etc.) can be represented in the steady state as Fourier series, and

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† Coles Signal Laboratory, Red Bank, N. J.

[1] Boyajian, "Theory of d.c. excited iron-core reactors and regulators," *Trans. AIEE*, vol. 43, pp. 119–936; June, 1924.

[2] W. Dällenback and E. Gerecke, "Die Strom- und Spannungsverhältnisse des Grossgleichrichters," *Archiv für Elek.*, vol. 14, pp. 171–248; January, 1925.

[3] Th. Buckhold, "On the theory of the magnetic amplifier," *Archiv für Eleck.*, vol. 37, No. 4, pp. 197–211; 1943.

[4] (a) A. Uno Lamm, "The Transductor, D.C. Pre-saturated Reactor, with Special Reference to Transductor-Control of Rectifier," Esselte Aktiebolag, Stockholm, 1943.

(b) A. Uno Lamm, "Some fundamentals of a theory of the transductor or magnetic amplifier," *Trans. AIEE*, vol. 66, pp. 1078–1085; April, 1947.

[5] H. S. Kirschbaum and E. L. Harder, "A balanced amplifier using biased saturable core reactors," *Trans. AIEE*, vol. 66, pp. 273–278; January, 1947.

(b) that the amplifier output can be obtained as a small variation of the state existing at the quiescent point. Starting with the work of Peterson[6] the first assumption can be shown analytically to be correct, and approximate expressions for the Fourier coefficients can actually be derived in terms of the hysteresis loop coefficients. The second assumption implies that this work is concerned only with the "small signal theory" of the amplifier.

II. Amplifying Element

In the barest essentials, the magnetic amplifier consists of a D-shaped or toroidal core on which two coils are wound; one of the coils is supplied with a constant source of alternating voltage and the other coil is supplied with a bias dc voltage, and with the signal voltage. Alone, this simple arrangement has limited practical utility because of the unavoidable mutual coupling between the two coils. The practical amplifier is obtained by using two of these simple units either assembled on the same core (i.e., 3-legged core construction) or on separate cores and connecting the winding in such a way that the interaction between the ac and control windings cancels. That is, if M_{pc} represents the mutual inductance between the two windings,

$$M_{pc} = M_{cp} = 0 \tag{1}$$

is the condition that must obtain for the practical amplifier.

III. Nomenclature and Notation

The circuit which is considered is illustrated in Fig. 1. Interwinding capacities and leakage inductances are

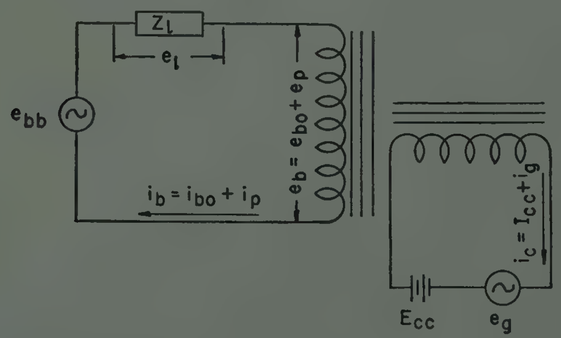

Fig. 1—Magnetic amplifier with load, symbolic representation.

neglected. In accordance with the previous remarks, the amplifying element is indicated in symbolic form. Quantities immediately related to the ac side of the amplifier will be referred to as *output* quantities and quantities related to the dc or control side as *input*, control, or signal quantities. From the circuit diagram, it appears that an ac generator of instantaneous voltage e_{bb} volts and angular frequency ω (both dependent on design) supplies the output coil consisting of N_p turns in series with a general load z_l. The voltage across the load is e_l and that across the output coil is e_b. The instantaneous current flowing in the output circuit under these conditions is i_b. The input coils are supplied with a voltage e_c which equals the sum of the dc bias voltage E_{cc} and the signal voltage e_g. The current flowing in this circuit is i_c.

It is convenient to split up these instantaneous currents and voltages as follows:

$$\begin{aligned} e_b &= e_{b0} + e_p \\ i_b &= i_{b0} + i_p \\ i_c &= I_{cc} + i_g \end{aligned} \tag{2}$$

where the quantities with the double subscript indicate steady values obtaining before the application of any varying voltages in the output or input circuit, and the quantities with the single subscript on the right-hand side of the (2) variations from these steady values.

The current i_b flowing in the input coil produces a magnetic field intensity h_b and the current i_c flowing in the input winding produces a magnetic field intensity h_c. The output coil voltage e_b and the control coil voltage e_c generate in the material a flux ϕ_b and ϕ_c, respectively. The material is assumed to have a static permeability μ and an incremental permeability μ_Δ. In general, in dealing with magnetic quantities the same subscript will be used as for the corresponding electric quantity. The length of the magnetic path is denoted by S_p and S_c for the ac flux and dc flux, respectively. The cross-sectional area is assumed to be the same for both paths and is denoted by A.

Absolute values are denoted by a bar over the symbol representing the quantity or by the conventional $| \, |$ sign enclosing the symbol. Vector quantities are represented by capital letters, except in the case of harmonic components where lower case letters are used.

IV. Steady-State Characteristics of Output and Input Circuits

A. Output Circuit

Consider the output circuit of the amplifier shown in Fig. 1. It consists of an ac generator supplying a load assumed to have a series resistance r_l and inductance l_l (assumed constant) in series with the output winding of the amplifier. This winding presents to the generator an inductance $l_p(t)$ and an effective resistance $r_p(t)$ made up of the actual resistance of the winding and of the resistance due to such effects as hysteresis, eddy current, and residual losses.[7]

Both the inductance and effective resistance are not constant, but vary in a complex way over the applied

[6] E. Peterson, "Harmonic production in ferro magnetic materials," *Bell Sys. Tech. Jour.* vol. 7, pp. 762–796; October, 1928.

[7] V. E. Legg, "Survey of magnetic materials and application in telephone system," *Bell Sys. Tech. Jour.*, vol. 18, pp. 438–664; July, 1939.

voltage cycle. This is made evident by the peculiar shape of the hysteresis loop and the variation of eddy current losses over this loop.[8] Moreover, the hysteresis and eddy-current components of $r_p(t)$ are a function[6,7] of the permeability μ.

The differential equation of this circuit can be written immediately

$$e_{bb} = (r_l + r_w)i_b + l_l \frac{di_b}{dt} + r_p'(t)i_b + l_p(t)\frac{di_b}{dt} \tag{3}$$

where r_w is the ohmic component of $r_p(t)$ and $r_p'(t) = r_p(t) - r_w$.

Assuming that the solution of (3) for the steady state yields for i_b a Fourier Series and using for convenience the complex form,[9] the current can be expressed as

$$i_b = \sum_{s=-\infty}^{+\infty} i_{bs}e^{js\omega t}. \tag{4}$$

Similarly, it is also convenient to assume the inductance $l_p(t)$ to be represented by a Fourier Series,

$$l_p(t) = \sum_{k=-\infty}^{+\infty} l_k e^{jk\omega t}$$

and also for $r_p'(t)$

$$r_p'(t) = \sum_{k=-\infty}^{+\infty} r_k e^{jk\omega t}. \tag{5}$$

Assuming, further, that the supply voltage is sinusoidal and given by,

$$e_{bb} = Re\overline{E}_{bb}e^{j(\omega t+\theta)} \tag{6}$$

equation (3) can be written by substituting (4), (5), and (6) and performing the differentiations indicated:

$$Re\overline{E}_{bb}e^{j(\omega t+\theta)} = \sum_{s=-\infty}^{+\infty}(r_l' + jx_{ls})i_{bs}e^{js\omega t} + \sum_{\substack{s=-\infty\\k=-\infty}}^{+\infty}(r_k' + j\omega s l_k)i_{bs}e^{j(s+k)\omega t} \tag{7}$$

where:

$$r_l' = r_l + r_w$$
$$x_{ls} = s\omega l_1.$$

The infinite double sum in (7) is the voltage across the output winding less the drop due to r_w. Denoting by e_{bn}' its n^{th} harmonic, it can be shown[10] to be given by

$$e_{bn}' = \sum_{k=-\infty}^{+\infty}[r_k' + j(n-k)\omega l_k]i_{b(n-k)}. \tag{8}$$

In general, then, the voltage e_{bn} is complex; accordingly, it will be assumed that:

$$e_{bn}' = \bar{e}_{bn}'e^{j\lambda_n}. \tag{9}$$

Substituting (9) in (7) and equating coefficients of the same harmonic on both sides of the equation, it obtains:

$$E_{bb}\delta_n = \bar{z}_{ln}'e^{j\theta'_{ln}} + \bar{e}_{bn}'e^{j\lambda_n} \tag{10}$$

where:

$$\bar{z}_{ln}'e^{j\theta'_{ln}} = r_l' + jx_{ln}$$

and[10]

$$\delta_n = 1 \text{ for } n = 1; \quad \delta_n = 0 \text{ for } n \neq 1.$$

Since all the quantities in equations (10) are sinusoidal quantities of the same frequency, the equation can be represented vectorially. The vector diagram is illustrated in Fig. 2. The magnitude of the voltage

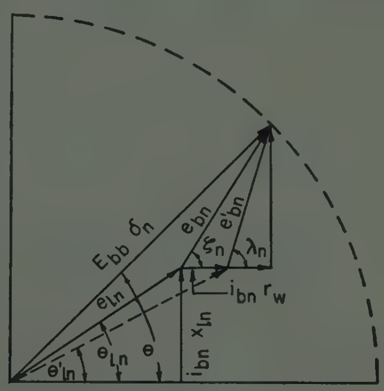

Fig. 2—Vector diagram of output circuit for the n^{th} harmonic under no signal condition.

$E_{bb}\delta_n$ is assumed constant. Therefore, the terminus of its vector is constrained to move in the circular path indicated.

B. Input Circuit

Similar considerations as above can be applied to the input circuit. However, for the purpose of the present discussion i_c will be assumed sinusoidal and to be given by:

$$i_c = ReI_ce^{jqt}. \tag{11}$$

V. Amplifier Characteristics

A. General Equation for i_b

The voltage e_b developed across the output circuit is a function of the state of saturation of the material forming the core. For a given material and design, the state of saturation will depend on the total instantaneous currents i_b and i_c and also on di_b/dt. The dependence on the time derivative of i_b is not immediately obvious. To show this, imagine that the output circuit of the ampli-

[8] "Magnetic Circuits and Transformers," MIT E. E. Staff, John Wiley & Sons, Inc., New York, N. Y., Chap. VI, 1943.
[9] E. A. Guillemin, "Communication Networks," John Wiley & Sons, New York, N. Y., vol. 1, Chap. X, 1931.
[10] See page 404 of footnote reference 9.

fier is equivalent to a variable resistance r_p in series with a variable inductance l_p. Then, in the absence of control coil excitation, e_b can be expressed as the voltage $f(i_b)$ across r_p and the voltage $d\lambda/dt$ across l_p. λ is the effective flux linkages. Thus:

$$e_b = f(i_b) + \frac{d\lambda}{dt}$$

which can also be written

$$e_b = f(i_b) + \frac{d\lambda}{di_b} i_b', \tag{12}$$

where:

$$i_b' = \frac{di_b}{dt}.$$

It follows from (12) that for small variations, the general expression for e_b in the presence of control coil excitation can be written:

$$e_b = F(i_b, i_b', i_c). \tag{13}$$

B. General Expression for i_p

Referring to equation (13) the variation of e_b can be written immediately (see (2))

$$e_p = \frac{\partial e_b}{\partial i_b} i_p + \frac{\partial e_b}{\partial i_b'} i_p' + \frac{\partial e_b}{\partial i_c} i_g \tag{14}$$

where:

$$i_p' = \delta(i_b').$$

In (13), as well as in the sequel, the partial coefficients will be understood to be evaluated at the quiescent point.

The variations e_p and i_p contains components that are products of the interaction of the signal frequency q and of the exciting voltage frequency ω. Thus it will be assumed that these variations have the form:

$$e_p = Re \sum_\nu b_\nu e^{j\beta_\nu t} \tag{15}$$

$$i_p = Re \sum_\nu a_\nu e^{j\beta_\nu t} \tag{16}$$

where β_ν includes frequencies $nw \pm lq$, ($n = 0, 1, 2, 3 \cdots$; $l = 1, 2, 3 \cdots$). The expression for i_p' may be obtained by differentiating (16);

$$i_p' = \delta i_b' = \frac{d}{dt}(\delta i_b) = Re \sum_\nu j\beta_\nu a_\nu e^{j\beta_\nu t}. \tag{17}$$

In (17), q may be neglected whenever it appears in the coefficients β_ν, except when $nw = 0$ since in practice $q << \omega$. Equation (12) gives for the first two partial coefficients of (14):

$$\frac{\partial e_b}{\partial i_b} = \frac{df(i_b)}{di_b} + \frac{d}{dt}\left(\frac{d\lambda}{di_b}\right) \tag{18}$$

$$\frac{\partial e_b}{\partial i_b'} = \frac{d\lambda}{di_b}. \tag{19}$$

Now, $df(i_b)/di_b$ and $d\lambda/di_b$ are functions of e_b and therefore may be expanded in Fourier series with angular frequency ω. The coefficient $d\lambda/di_b$ will be recognized as the inductance of the output winding. Its Fourier expression may be written as in (5):

$$\frac{\partial e_b}{\partial i_b} = \frac{d\lambda}{di_b} = \sum_{k=-\infty}^{+\infty} l_k e^{jk\omega t}. \tag{20}$$

If the expansion for $df(i_b)/di_b$ which will be recognized as the loss resistance r_p, is assumed to be given by

$$\frac{df(i_b)}{di_b} = \sum_{k=-\infty}^{+\infty} r_k e^{jk\omega t} \tag{21}$$

then, from (18), (20), and (21)

$$\frac{\partial e_b}{\partial i_b} = \sum_{k=-\infty}^{+\infty} (r_k + jk\omega l_k) e^{jk\omega t}. \tag{22}$$

Considering now the partial coefficient $\partial e_b/\partial i_c$, which obviously is the dynamic mutual impedance between input and output circuits, heuristically it would be expected that this coefficient is a function of both ω and q. Therefore, it will be represented by the trigonometric series,

$$\frac{\partial e_b}{\partial i_c} = Re \sum_\nu z_{m\nu} e^{j\beta_\nu t}. \tag{23}$$

Substituting (23), (22), and (20) together with (17), (16), (15), and (11) into (14), and equating coefficients of the same β_ν linear relations are found among the coefficients, which can be written in expanded matrix form thus (note that $b_q = a_q = 0$ by equation (1))

$$\begin{pmatrix} \vdots \\ b_{q+\omega} \\ 0 \\ b_{q-\omega} \\ \vdots \end{pmatrix} = \begin{pmatrix} & \vdots & \vdots & \vdots & \\ \cdots & z_{+10} & z_{+1+1} & z_{+1+2} & \cdots \\ \cdots & z_{0-1} & z_{00} & z_{0+1} & \cdots \\ \cdots & z_{-1-2} & z_{-1-1} & z_{-10} & \cdots \\ & \vdots & \vdots & \vdots & \end{pmatrix} \begin{pmatrix} \vdots \\ a_{q+\omega} \\ 0 \\ a_{q-\omega} \\ \vdots \end{pmatrix}$$

$$+ \begin{pmatrix} \vdots & \vdots & \vdots \\ z_{m+10} & z_{m+1+1} & z_{m+1+2} \\ z_{m00} & z_{m0+1} & z_{m0+2} \\ z_{m-10} & z_{m-1+1} & z_{m-1+2} \\ \vdots & \vdots & \vdots \end{pmatrix} \begin{pmatrix} i_{g+1} \\ 0 \\ i_{g+1} \end{pmatrix} \tag{24}$$

where:

$$z_{kn} = r_n + j(k\omega + q)l_n$$
$$z_{mk_n} = z_{m(k\omega+nq)}. \tag{25}$$

The column vector $b(\equiv e_p)$ in equation (24) can be expressed in terms of the a_p's$(\equiv i_p$'s$)$. Referring to Fig. 2, the voltage e_{bn} can be written:

$$e_{bn} = r_w i_{b_n} + e_{b_n}'. \tag{26}$$

Next, consider that equation (10) can be shown to apply here because the changes involved are small, and $q << \omega$ in practice. The voltage variation e_{pn}' can be obtained by taking the variation of (10). The derivation is straight forward but rather lengthy, and will not be given here. The result obtained, assuming that θ'_{ln}, λ_n and $\overline{E}_{bb}$ are constant, is:

$$e_{pn} = -z_{ln}\Theta i_{pn} \tag{27}$$

where:

$$\Theta = \frac{[1 - \delta_n e^{j(\pi/2-\theta_{ln}'+\theta)} \sin(\theta_{ln}' - \theta)][1 - \delta_n e^{j(\lambda_n-\theta-(\pi/2))} \sin(\lambda_n - \theta)]}{1 - \delta_n \sin^2(\lambda_n - \theta)}. \tag{28}$$

It will be noted that $\Theta = 1$ in all cases in which $n \neq 1$. From (26) and (27) the variation e_{pn} is therefore

$$e_{pn} = [r_w - z_{ln}\Theta]i_{pn} \equiv -\zeta i_{pn}. \tag{29}$$

It follows from (29) that the vector $\boldsymbol{e}_p$ can be written as the product of a diagonal matrix and column vector $\boldsymbol{i}_p$ thus

$$\boldsymbol{e}_p = -\boldsymbol{Z}_l \boldsymbol{i}_p \tag{30}$$

where:

$$\boldsymbol{Z}_l = \begin{bmatrix} \ddots & & & & \\ & \zeta_{+1} & & & \\ & & \zeta_0 & & \\ & & & \zeta_{-1} & \\ & & & & \ddots \end{bmatrix}. \tag{31}$$

Inserting (30) in (24) and using abridged matrix notation throughout to save space, (24) takes the form:

$$[\boldsymbol{Z}_p + \boldsymbol{Z}_l]\boldsymbol{i}_p = -\boldsymbol{Z}_m \boldsymbol{i}_g. \tag{32}$$

Equation (32) is the fundamental equation of the amplifier. In it, Z_l is a diagonal matrix (31), but Z_p and Z_m are not. Immediately the question arises whether these matrices can be reduced to true diagonal form. Discussion of this question is beyond the scope of this paper. However, in order to simplify the following discussion, it will be assumed here that diagonalization has been affected. Then, it can be written generally

$$(z_{p_n} + \zeta_n)i_{p_n} = -z_{m_n}i_{g_n} \tag{33}$$

where z_{p_n} and z_{m_n} are characteristic values[11] of $\boldsymbol{Z}_p$ and $\boldsymbol{Z}_m$. It follows that in the case in which i_g is sinusoidal only the first two sidebands of i_{p_n} are present (i.e., $\omega+q; -\omega+q$). In the sequel, consideration will be limited to the last case. No confusion will arise if the sideband subscript is dropped and equation (33) written simply:

$$(z_p + \zeta)i_p = -z_m i_g. \tag{33a}$$

From equations (33), (32) or (33a), the equivalent circuit of the amplifier is easily visualized. It will be noted that it possesses a time constant which depends on the internal characteristics and on the character of the load and not only on the impedance of the output circuit.[12]

D. Conditions for Maximum Amplification

Consider equation (33a). It will be apparent that, for a given set of operating conditions, there must exist some value of load reactance x_l between minus infinity and infinity and some value of load resistance r_l' between zero and infinity at which maximum gain will be obtained. This follows because at the limits indicated for these parameters the gain is zero.

(1) Optimum load reactance.

The current gain of the amplifier will be defined by:

$$G = \frac{-z_m}{z_p + \zeta}. \tag{34}$$

Then, the equation

$$\frac{dG}{dx_l} = \frac{d\zeta}{dx_l} = 0, \tag{35}$$

when solved for x_l, should yield the optimum value of the reactive component of the load.

Obtaining the derivative in (35) by differentiating (29) and (28), after factorization and collection of real and imaginary terms the result is:

$$P\frac{d\bar{z}_l}{dx_l} + \bar{z}_1\frac{dP}{dx_l} + j\left(Q\frac{d\bar{z}_l}{dx_l} + \bar{z}_1\frac{dQ}{dx_l}\right) \tag{36}$$

where:

[11] H. Margenau and G. M. Murphy, "The Mathematics of Physics and Chemistry," p. 301; D. Van Nostrand Co., Inc., New York, N. Y., 1943.

[12] See Section VIII of footnote reference 4b.

$$\begin{aligned} P &= 1 - \sin^2\psi + \sin^2\tau - \sin\tau\sin\psi\cos(\psi-\tau) \\ Q &= \tfrac{1}{2}\sin 2\psi - \tfrac{1}{2}\sin 2\tau + \sin\tau\sin\psi\sin(\psi-\tau) \end{aligned} \tag{37}$$

and

$$\psi = \lambda + \theta_l' - \theta$$

$$\tau = 2\theta_l' - \theta.$$

In order for equation (36) to be satisfied, both the real and imaginary components must vanish. Considering, for instance, the real component, taking z_1 defined by (10) and noting that

$$\frac{dP}{dx_l} = \frac{dP}{d\theta_l'}\frac{d\theta_l'}{dx_l} = \frac{dP}{dx_l}\frac{\cos^2\theta_l'}{r_l'},$$

it obtains for the optimum reactance x_{0p}

$$x_{0p} = -r_l'\left[\frac{P}{2\cos^2\theta_l'}\frac{1}{\dfrac{dP}{d\theta_l'}} \pm \frac{1}{2}\sqrt{\frac{P^2}{\cos^4\theta_l'}\frac{1}{\left(\dfrac{dP}{d\theta_l'}\right)^2} - 4}\,\right]. \tag{38}$$

It appears from equation (46) that, since x_{0p} must be real, it must be negative; that is, the load reactance for maximum gain must be capacitive. Buckhold[3] in his work arrives at a similar conclusion. Also, it is believed, that this might explain the experimental results of Kirschbaum and Harder[5] who obtained a higher value of amplification when they employed second and fourth harmonic filters in the output. These filters would appear capacitive to the fundamental.

(2) Optimum load resistance.

The load resistance for maximum power amplification will be determined for the case in which the load reactance is zero (i.e., $\theta_l'=0$) because in the general case the equations become too unwieldy.

The power amplification A will be defined by

$$A = \frac{r_l}{r_c}|G|^2 \tag{39}$$

and, therefore, the condition for maximum power amplification

$$\frac{dA}{dr_l} = 0 \tag{40}$$

is equivalent to

$$|\zeta + z_p|^2 - 2r_l|\zeta + z_p|\frac{d|\zeta + z_p|}{dr_l} = 0. \tag{41}$$

Writing:

$$z_p = r_w + r_{he} + jx_p$$

where:

r_{he} = the hysteresis, eddy currents, and residual loss component, and making the following assumptions: The r_w term in r_l' can be neglected in comparison with r_l and the $r_l'(r_w+r_{he})$ terms are negligible in comparison with r_l^2 terms, the result obtained from (41) for the value of optimum resistance r_{0p} is

$$r_{0p} = [1 - \sin(\lambda - \theta)]\sqrt{\frac{\bar{z}_p^2 - r_w^2 - 2r_w r_{he}}{P^2 + Q^2}} \tag{42}$$

where P and Q are as defined in (37) with the added condition that $\theta_l'=0$.

Equation (42) is not easily interpreted, because in effect it is not in closed form. However, for the fundamental component, if the assumption is made that $\lambda=\pi/2$ and r_w is negligible (42) reduces to:

$$r_{0p} = \sqrt{x_{p0}x_p}, \tag{43}$$

where x_{p0} is the reactance of the output winding at the operating point. Applying (43) to the experimental data of Kirschbuam and Harder[5] it is found that the amplification for the value of resistance computed from (43) is within about 1 per cent of the experimental value. The value of resistance computed from (43) is too high (i.e., about 8 per cent). This would, however, be expected on the basis of the assumptions made.

VI. Amplifier Characteristics in Terms of Magnetic Quantities

A. Expression for Amplification

Additional insight into the operation of the magnetic amplifier is obtained by expressing the amplification in terms of the magnetic field intensity, the flux density, and the permeability. This can be done in the general case; but it will, here, serve the purpose better to consider equation (33a) under the assumptions that:

(1) losses are absent (i.e., $r_w=0; \lambda=\pi/2$)

(2) the load is a pure resistance (i.e., $\theta_l'=0$).

Then, substituting for ζ under these conditions from equation (29), it obtains for i_p.

$$i_p = \frac{-x_m}{x_p + \dfrac{r_l\cos\theta\sin\theta}{1-\sin^2\theta}}\,i_g. \tag{44}$$

It can be shown that (44) can be written

$$i_p = \frac{-\dfrac{\partial e_b}{\partial i_c}}{\dfrac{\partial e_b}{\partial i_b} + x_{p0}}\,i_g, \tag{44a}$$

where the partial coefficients have been written in place of x_m and x_p and x_{p0} is the reactance of the output winding at the operating point.

Considering, now, the equalities

$$-{}_t e_b = N_p A \frac{db_b}{dt} = \frac{d}{dt} f(h_b, h_c)$$

$$= 4\pi \frac{d}{dt} f\left(\frac{N_p}{S_p} i_b, \frac{N_c}{S_c} i_c\right)$$

and

$$L_{p0} = \frac{4\pi N_p^2 A \mu_{\Delta 0} 10^{-7}}{S_p}$$

where $\mu_{\Delta 0}$ and L_{p0} are, respectively, the incremental permeability and inductance at the operating point, it is easily shown that (44a) takes the form

$$i_p = -\frac{N_c S_p}{N_p S_c} \frac{\dfrac{\partial b_b}{\partial h_c}}{\dfrac{\partial b_b}{\partial h_b} + \mu_{\Delta 0}} i_g. \qquad (45)$$

Equation (45) is the desired equation. It expresses the change in output current in terms of the design of the core and winding and in terms of the fundamental characteristics of the core material.

B. Criteria for Maximum Amplification

The partial coefficients appearing in (45) can be interpreted in terms of permeabilities by reference to Figs. 3 and 4 which illustrate, in greatly exaggerated form, displaced loop under conditions of constant h_b and

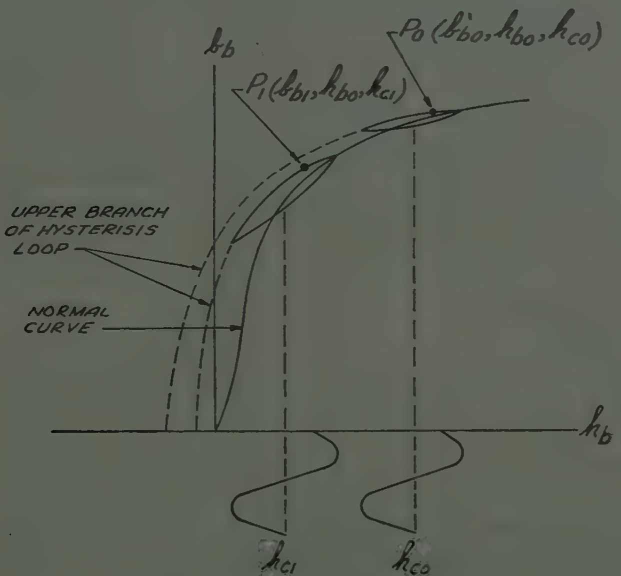

Fig. 3—Displaced hysterisis loops, idealized, under conditions of constant h_b and at two values of polarizing magnetic intensity h_c.

h_c, respectively. This interpretation is based on the assumption that, for small changes, the following relation holds approximately.

$$b_b \cong \mu h_c + \mu_\Delta h_b. \qquad (46)$$

Consider Fig. 3. Forming the quotient of the change in flux to the change in h_c in passing from point P_0 to point P_1, in terms of equation (46), and passing to the limit, it is found that

$$\frac{\partial b_b}{\partial h_c} \cong \mu_0 - |h_{c0}| \frac{\partial \mu}{\partial |h_c|} - |h_{b0}| \frac{\partial \mu_\Delta}{\partial |h_c|}. \qquad (47)$$

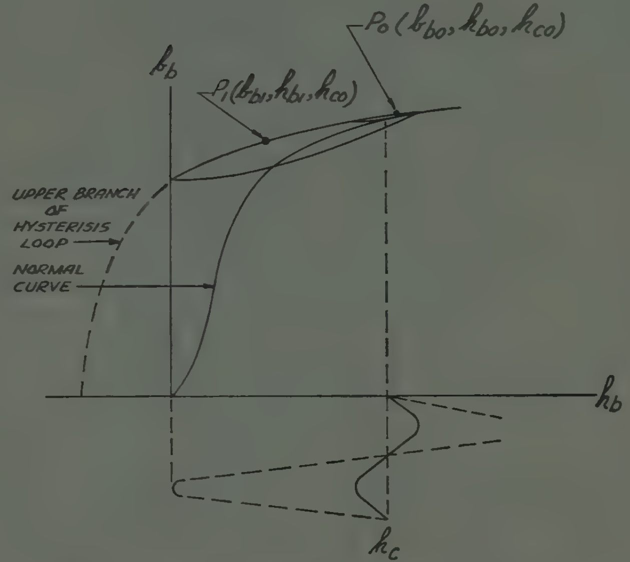

Fig. 4—Displaced hysterisis loops, idealized, under conditions of constant h_c and variable h_b.

In a similar way, considering Fig. 4, it is found that

$$\frac{\partial b_b}{\partial h_b} \cong \mu_{\Delta 0} + |h_{b0}| \frac{\partial \mu_\Delta}{\partial |h_b|}. \qquad (48)$$

Substituting (47) and (48) into (45) the desired expression is

$$i_p \cong -\frac{N_c S_p}{N_p S_c} \frac{\mu_0 - |h_{c0}| \dfrac{\partial \mu}{\partial |h_c|} - |h_{b0}| \dfrac{\partial \mu_\Delta}{\partial |h_c|}}{2\mu_{\Delta 0} - |h_{b0}| \dfrac{\partial \mu_\Delta}{\partial |h_b|}} i_g. \qquad (49)$$

Equation (49) gives interesting insight regarding the conditions to be met and those characteristics which a magnetic material should possess to make the best possible amplifier. For a given winding and core design, the amplification will be the larger, the more pronounced is the following inequality:

$$\mu_0 - |h_{c0}| \frac{\partial \mu}{\partial |h_c|} - |h_{b0}| \frac{\partial \mu_\Delta}{\partial |h_c|} >> 2\mu_{\Delta 0} - |h_{b0}| \frac{\partial \mu_\Delta}{\partial |h_b|}. \qquad (50)$$

It would, now, be very important to determine even qualitatively, whether practical magnetic materials actually possess characteristics which satisfy the inequality predicted by the theory.

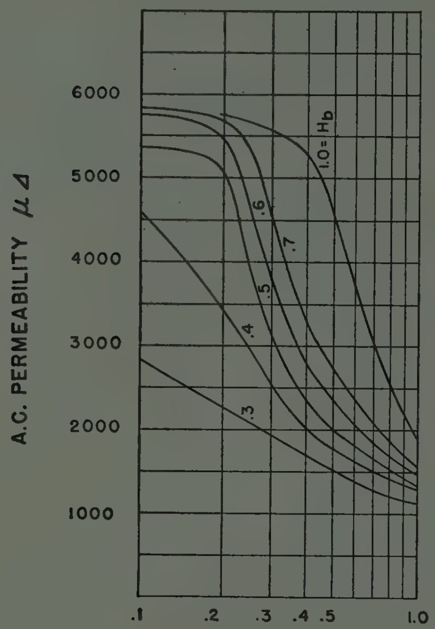

Fig. 5—Polarizing magnetic intensity H_{dc}-oersteds.

Data which would permit an extensive and direct check of the expression (50) is not available, however, for at least one material—U.SS Transf. 72, 29-gauge data was found[13] which could be put into the form to permit such check. These data are presented in Figs. 5 and 6. A perusal of these figures will show that (50) is satisfied in the region of the characteristics where

$$\frac{\partial \mu_\Delta}{\partial |h_c|} << 0$$

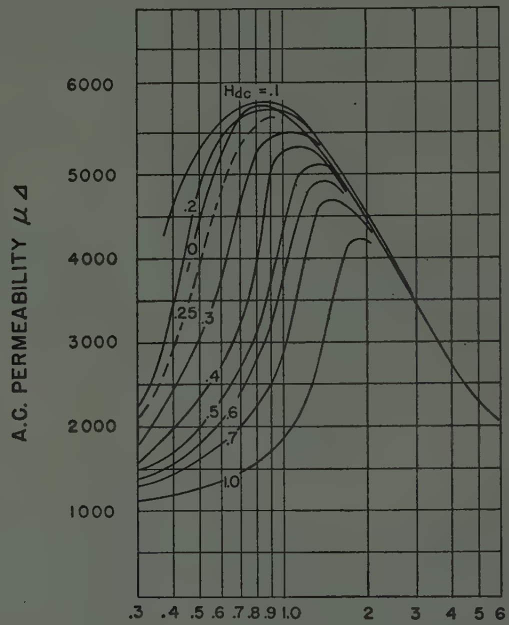

Fig. 6—Alternating-current magnetic intensity H_b-oersteds.

and

$$\frac{\partial \mu_\Delta}{\partial |h_b|} \cong \frac{\partial \mu}{\partial |h_c|} > 0.$$

[13] S. S. Attwood, "Electric and Magnetic Fields," John Wiley & Sons, New York, N. Y., p. 334; 1941.

VII. Acknowledgment

The author wishes to express his appreciation for the encouragement and facilitation given by J. Hessel, Chief, Radio Communication Branch, Coles Signal Laboratory, and R. E. Lacy, Chief, Radio Relay and Microwave Section, Coles Signal Laboratory, without which this paper would not be possible. Also appreciation is extended to Georg Goubau of Coles Signal Laboratory and E. Both of Squier Signal Laboratory, for their interest and valuable criticism.

Correction

Leo Storch, author of the paper, "Design Procedures for Pi-Network Antenna Couplers," which appeared on pages 1427–1432, of the December, 1949, issue of the Proceedings of the I.R.E., has brought the following errors to the attention of the editors:

In Fig. 2, page 1429, the centers of the two circles should be labeled $(-X_3')$ and $(-X_3'')$, respectively.

In the biography of Mr. Storch on page 1443, the advanced degree should read M. S., instead of M. A.

Receiving Tubes Employing Secondary Electron Emitting Surfaces Exposed to the Evaporation from Oxide Cathodes*

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Summary—Previous use of secondary-emission multiplication in receiving tubes has been accompanied by difficulties with life and has been accomplished in oxide cathode tubes by shielding the secondary emitter from cathode evaporation. This shielding required a complicated tube structure and additional electrodes and leads. This paper describes methods of controlling cathode evaporation to obtain satisfactory tube life and even to enhance secondary emission gain in simple tube constructions in which the secondary emission surface is directly exposed to the oxide cathode evaporation. A 1.4-volt filament tube is described which may be operated to give a transconductance of three times the normal input transconductance, or may be operated to give the normal input transconductance with a reduction to 40 per cent of the original total battery power. The construction and characteristic curves of an indirectly heated oxide cathode tube with the dynode directly exposed to the cathode are given. A transconductance of 24 ma/volt and a wide-band figure of merit of two to three times that of conventional tubes is obtained. A high cathode temperature reduces dynode life, but a variation of ±10 per cent in heater voltage seems satisfactory.

Introduction

THE IDEA OF applying secondary electron emission as a means of amplification has been appealing for many years.[1] This form of multiplication or amplification has been used for about a decade in photoelectric multipliers[2] but has been used to only a very limited extent in grid controlled tubes.[3,4] A satisfactory life in the photoelectric multiplier is easier to obtain because of the low current densities involved and the absence of the contamination produced by an oxide cathode.

The present interest in wide-band amplifiers has led to a reconsideration of the application of secondary electron emission multipliers to conventional grid-input structures. Thompson[5] explained some of the factors involved, and a summary of the characteristics that can be obtained by secondary emission amplification leads to the following:

(1) For wide-band amplification, it is well known that a figure of merit for a tube is proportional to the ratio of transconductance to capacitance. By applying secondary emission to a conventional grid structure the transconductance is multiplied by the secondary-emission amplification without increasing capacitance.

(2) The input conductance is left unchanged.

(3) The amplification of the added multiplier can be made essentially frequency independent up to frequencies of the order of a thousand megacycles.

(4) The additional noise introduced is small.

(5) One can obtain a high transconductance without an extremely critical input structure. Although one substitutes the possible difficulty of a critical secondary-emission (dynode) surface, our experience indicates that the control of this surface is readily possible with present receiving-tube manufacturing techniques.

Early experimenters with secondary electron multiplication soon found that the secondary emission yield of a surface decayed very rapidly on life when exposed to an oxide cathode, and sought to overcome these difficulties by shielding the secondary emitter or dynode from cathode evaporation.[3,4] Even with this shielding, considerable difficulty with the life of the secondary emitter was experienced. Secondary emission amplification has also been applied to laboratory beam-deflection amplifier tubes[6] and, since the war, much work has been done in these laboratories to use secondary emission in other tube types.

Contrary to previous results, in which exposure to oxide cathodes resulted in rapid decay of secondary emission, experiments to be discussed will show that cathode evaporation difficulties can be overcome and such evaporation may even be used to enhance secondary emission. A small amount of evaporation can be advantageous; a large amount may be disastrous.

I. Experimental Methods

Fig. 1 shows a type of structure used for many of the secondary emission measurements. It consists of a low voltage gun (adapted from the beam-deflection tube)[7] giving a rectangular beam with a current density of 50 to 75 ma per square cm, a set of deflection plates to permit one dimensional scanning, a series of apertures to eliminate stray currents, a collector, and a rotating

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† RCA Laboratories, Princeton, N. J.

[1] V. K. Zworykin, G. A. Morton, and L. Malter, "The secondary emission multiplier—a new electronic device," Proc. I.R.E., vol. 24, pp. 351–375; March, 1936.

[2] V. K. Zworykin and J. A. Rajchman, "The electrostatic electron multiplier," Proc. I.R.E., vol. 27, pp. 558–566; September, 1939.

[3] J. L. H. Jonker and A. J. Overbeck, "Application of secondary emission in amplifying valves," *Wireless Eng.*, vol. 15, pp. 150–156; 1938.

[4] H. M. Wagner and W. R. Ferris, "The orbital-beam secondary-electron multiplier for ultra-high frequency amplification," Proc. I.R.E., vol. 29, pp. 598–602; November, 1941.

[5] B. J. Thompson, "Voltage-controlled electron multipliers," Proc. I.R.E., vol. 29, pp. 583–587; November, 1941.

[6] G. R. Kilgore, "Beam deflection control for amplifier tubes," *RCA Rev.*, vol. VIII, pp. 480–505; September, 1947.

[7] E. W. Herold and C. W. Mueller, "Beam deflection mixer tubes for u-h-f," *Electronics*, vol. 22, pp. 76–80; May, 1949.

dynode structure. Eight samples, which may all be different, are mounted on the dynode structure and, by rotating the tube assembly, any one of the dynodes may be brought in front of the collector for observation. By means of evaporators, which may be either filamentary or indirectly-heated cathodes in the corners of the collector, the effects of evaporation on the dynode may be studied. Although the emitter for the gun of this tube has an oxide cathode, the presence of any evaporated material from this cathode on the dynode has never been detected. Thus the gun structure shows that one may eliminate cathode contamination by the use of small apertures; two 0.004-inch apertures about $\frac{3}{8}$ inch apart so limit the solid angle that evaporation is negligible. Fig. 2 shows a photograph of this experimental tube. The glass structure is made long to keep the glass seals as far away from the test surfaces as possible and thus eliminate contamination from the heated glass during sealing in.

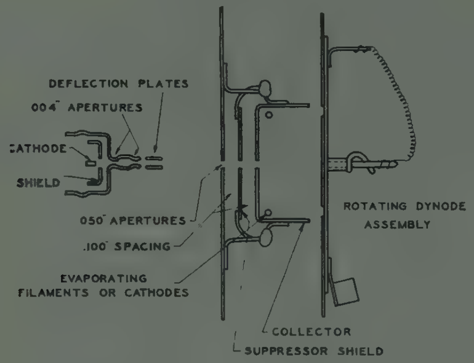

Fig. 1—Sectional view of tube for testing effect of evaporation on secondary emitter surfaces.

In this type of tube many dynode surfaces have been examined in an attempt to obtain most of the following desirable features:

(1) The secondary electron yield should be constant throughout life under high primary current density bombardment.

(2) The yield should be as high as possible at as low a voltage as possible.

(3) No processing should be necessary in the tube in which the emitter is to be applied, i.e., the secondary emitter should be capable of complete processing prior to use in a tube and thereafter no harm should come from ordinary handling, baking, and degassing procedures.

(4) Processing should be simple and easily reproduced.

(5) A long life should be obtained with direct exposure to oxide cathode evaporation.

(6) The dynode material should be nonmagnetic, capable of being punched, formed, and resistance-welded, and should not flake or scratch easily.

Although many materials have been studied with the above objectives in mind, the best results were obtained with processed silver-magnesium alloy,[8] about 2 per cent magnesium in silver. This alloy, after polishing, forming, and cleaning, is heated in a vacuum with a residual air pressure of 10^{-4} to 10^{-5} mm of mercury in order to diffuse magnesium to the surface and oxidize it. One then obtains a thin layer of magnesium oxide, which may contain some free magnesium or oxygen, on a silver-magnesium alloy base. This material is then used in the tube where no further treatment except outgassing is applied. In all the following discussions in this paper, unless it is specifically stated otherwise, the dynode used consists of this processed silver-magnesium alloy.

Fig. 3 shows the influence of the evaporated material from a regular 1.4-volt, 0.050-ampere, oxide-coated filament on preprocessed silver-magnesium alloy. The filament consists of a very small diameter nickel-tungsten alloy wire coated with a triple-carbonate spray. The thirty-second heating at 40 ma hardly changes the secondary-emission ratio, while each additional heating in steps of 10 ma increases the ratio, but the final heating at 85 ma does not change the ratio from the 70 ma heating. This dynode activation occurs at heating values at which the filament is also activated for thermionic emission, which is very convenient for the use of sec-

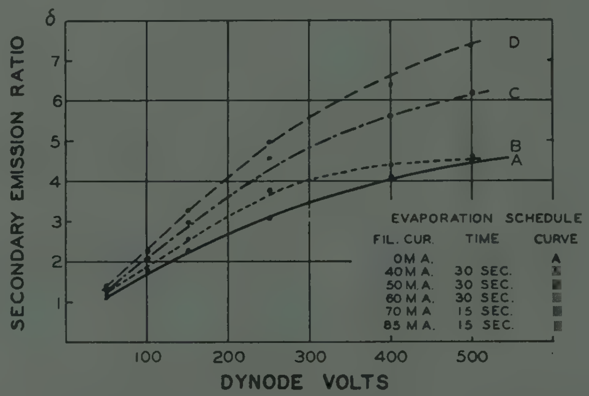

Fig. 3—Evaporation test of 1.4-volt filament on preprocessed silver-magnesium alloy.

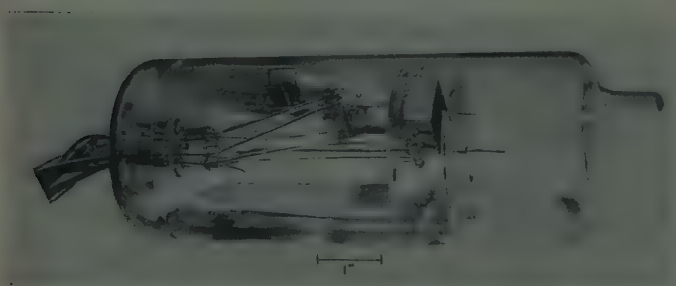

Fig. 2—Photograph of tube for testing secondary emitter surfaces.

[8] V. K. Zworykin, J. E. Ruedy and E. W. Pike, "Silver-magnesium alloy as a secondary emitting material," *Jour. App. Phys.*, vol. 12, pp. 696–698; September, 1941.

ondary-emission multiplication in filament tubes. Obviously, an activation which gave good secondary emission, but no primary emission would be inconvenient.

In the case of indirectly-heated cathodes, much larger volumes of coating material are available for evaporation and the problem of evaporation control is more difficult. However, experiments in the tube of Fig. 1 showed that control of evaporation to maintain life and also to enhance the secondary emission ratio was possible. Further data on the indirectly heated cathode will be illustrated by application to the tubes shown in Part III of this paper.

II. Secondary Emission in Filament-Type Tubes

Fig. 4 shows the construction of a developmental tube in schematic which is similar to that of the RCA Type

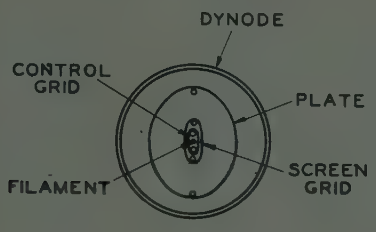

Fig. 4—Experimental filament-type secondary emission tube.

1U4 with the suppressor used as the plate. Primary electrons from the filament are accelerated by the screen grid, pass between the wires of the plate, and liberate secondary electrons at the dynode, which are then collected by the plate. The circuit of Fig. 5 shows the connections of the tube to a battery. Note that the

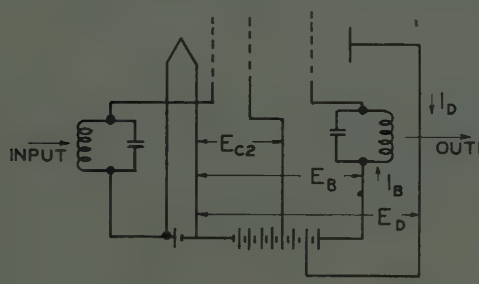

Fig. 5—Circuit of filament-type secondary emission tube.

current arising from the secondary electrons flows in plate-dynode loop and through only a section of the battery. This fact allows power saving possibilities as will be illustrated later. The output may be taken from the plate, in which case the full gain is realized; or, alternatively, the output may also be taken from the dynode, in which case the gain becomes decreased by unity. With dynode output, an amplifier is obtained with no phase reversal between input and output.

Fig. 6 shows results obtained in actual miniature tubes. Since, in this type of structure, some primary current is intercepted by the plate, a correction would be necessary to get the true secondary emission ratio; consequently, the secondary emission gain plotted is defined as I_B/I_B+I_D. This factor involves the efficiency of the output structure and is the important one in tube design. The tubes of lot A and C were given a factory Sealex automatic exhaust. Curve A gives the results of the combination of evaporation from an oxide-coated filament and a nickel-plated steel dynode which gives a gain less than unity and shows that the evaporated material is not thick enough to give a high-gain characteristic independent of the base metal. Curve B illustrates the performance of the processed silver-magnesium dynode with a tungsten filament and no evaporation. Curve C shows that the conditions for high gain are the combination of the evaporation from the oxide-coated filament and the preprocessed silver-magnesium dynode.

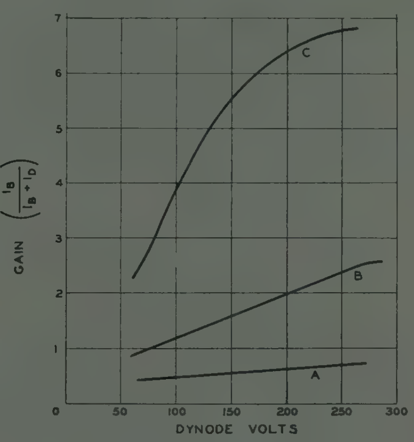

Fig. 6—Curves of secondary-emission gain versus dynode voltage: A—Oxide-coated filament and nickel-plated steel dynode. B—Tungsten filament and preprocessed silver-magnesium alloy dynode. C—Oxide-coated filament and preprocessed silver-magnesium alloy dynode.

In a battery-operated tube, one strives for a uniform high gain during life at low bombarding voltages. With a low bombarding voltage, only the first few atomic layers of the dynode surface are used and the dynode is particularly susceptible to the influence of evaporated material. The evaporation during life from an oxide-coated filament of very small diameter is low because of the relatively small quantity of oxide used. The ratio of dynode area to cathode area is very large, causing deposits on the dynode per unit area to be small. This large ratio of areas is an important feature of this type of tube construction. The low current density of $\frac{1}{4}$ to $\frac{1}{2}$ ma per square centimeter at the dynode does not impose stringent life requirements, and satisfactory life of over 500 hours has been obtained.

Table I gives typical performance data of a developmental filament tube, and also illustrates the power saving possibilities. The tube may be operated under conditions which give maximum gain or minimum power consumption. If the voltages on the secondary emitter tube are adjusted to give the same transconductance as

a standard tube, the 1U4 for instance, only 40 per cent of the 1U4 total power is required and even then 70 per cent of the remaining power is used for heating the filament. If the same power is used, twice the transconductance can be obtained. Other combinations are possible as shown in the Table and, though not shown, a transconductance of 5,000 μmho may be obtained at a dynode voltage of 250.

TABLE I

Typical Operating Conditions and Power Consumption of Secondary-Emission Filament-Type Tube Compared with Regular Tube

Tube	Transconductance μmho	Total Power* Watts	Applied Volts Control grid=0		
			Plate	Dynode	Screen
Developmental Secondary-Emission Type	900	0.10	90	67½	32
"	1,400	0.15	90	67½	45
"	1,800	0.27	90	67½	67½
"	2,300	0.34	112½	90	67½
"	2,700	0.46	135	90	67½
Standard 1U4	900	0.26	90	—	90
"	730	0.16	67½	—	67½

* Filament power of 0.07 watts included.

III. Application of Secondary-Emission Multiplication to Indirectly Heated Oxide Cathode Tubes

The application of secondary emitters directly exposed to an indirectly heated oxide-coated cathode is desirable because of the simplification obtained by eliminating some focussing electrodes and voltages inherent in the orbital beam structure.[3,4] The problem of operating the dynode directly exposed to the oxide-coated cathode in this case is much more difficult because of the large cathode used.

Fig. 7 shows the construction of an indirectly heated oxide-coated cathode tube. The structure, placed in a 7-pin miniature envelope, is similar to that of the standard 6AG5 with the plate replaced by a multiplier section. The electron stream as shown in Fig. 7, is directed between the plates to the dynode surface.

The plate-dynode configuration was arrived at by experiments on a rubber-membrane large-scale model, and then refined by experiment in tubes. Since the collection problem is critical and always exists in a secondary-emission tube, the features desired in the output section are listed below:

(1) A good collection field at low dynode-to-plate voltage differences even with high current densities. A low collection voltage is desirable to reduce plate heat dissipation requirements.

(2) Interception of a minimum of primary current by the collector or plate.

(3) Adequate dynode and plate heat dissipation.

(4) A small transit angle spread of the electrons.

(5) A low dynode current density.

(6) An arrangement which will accommodate large electrode voltage variations with a minimum change of tube characteristics.

A reasonable compromise between these requirements must be worked out and the over-all result obtained is illustrated in Figs. 8 to 10 showing characteristics of a developmental tube.

Many types of grid or input structure may be used with the secondary emission output and tests were actually made on two types. One was that of the standard RCA type 6AG5 tube which is a typical low-cost high-production electrode system. The other experimental type had closer spacings and gave an input transconductance, (i.e., transconductance measured as a normal pentode) of 7.5 ma per volt. This structure also had a higher transconductance-to-current ratio which is desirable in order to minimize direct-current losses in the tube and associated bleeders or power supplies. The performance of both experimental types was similar in that the output transconductance was about three times the input transconductance. The over-all tube characteristics are illustrated by characteristic curves from the tube having this higher transconductance input structure.

Fig. 8 shows the plate current family. Note the flat plate-current characteristic over a wide plate-voltage range. The plate current finally drops off due to the plate "robbing" current from the dynode so that this current is no longer multiplied by the secondary emis-

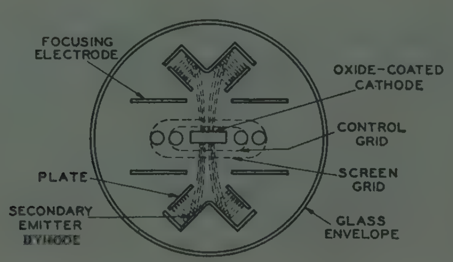

Fig. 7—Experimental high transconductance secondary emitter tube.

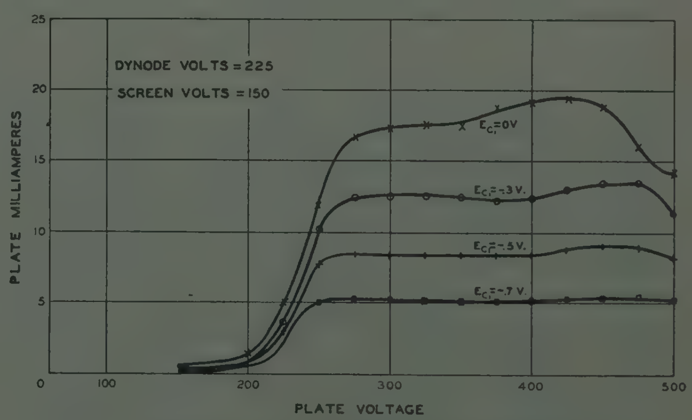

Fig. 8—Plate-current plate-voltage characteristics of tube of Fig. 7.

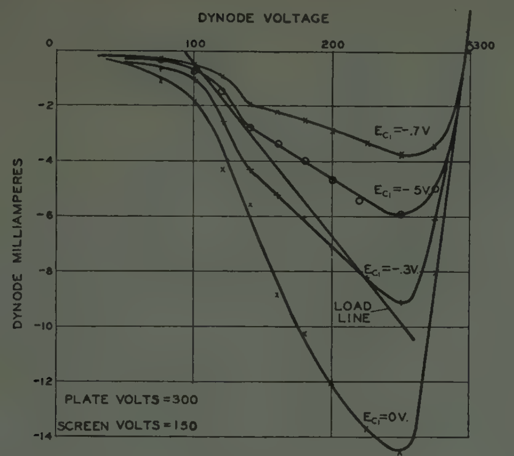

Fig. 9—Dynode-current dynode-voltage characteristics of tube of Fig. 7.

sion ratio. Fig. 9 shows the dynode current family, and a load line showing the maximum allowable resistance in the dynode circuit in order to insure stability as will be discussed later. This resistance may be an output load or the resistance of the power supply or network supplying the dynode voltage. Fig. 10 shows grid-control

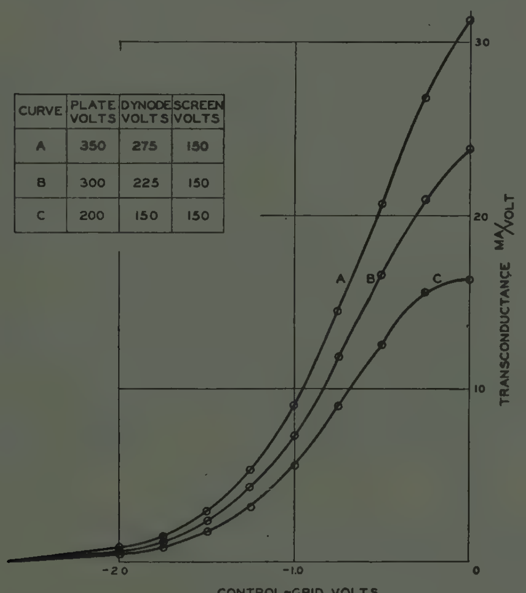

Fig. 10—Transconductance—control-grid voltage curves of the tube of Fig. 7.

characteristics plotted in terms of ma per volt, which gives a more convenient number for high transconductance tubes; multiplication by 1,000 will convert the units to micromhos if desired. Table II gives data comparing various tubes for wide-band operation and shows how a wide-band figure of merit of $2\frac{1}{2}$ times that of the type 6AK5 tube may be obtained. The close-spaced 404A[9] uses special equipment and techniques in its construction which are not normally used at present in receiving tube manufacturing technique.

TABLE II

Comparison of High-Transconductance Tubes

Description	*Transconductance* ma/v	*Equivalent Noise Resistance** Ohms	*Wide-Band Figure of Merit* $\frac{gm}{2\pi C \text{ total}}$ *Megacycles*
6AG5	5.0	1,600	72
6AK5	5.1	1,700	90
Multiplier type	24.0	1,300	250
404A	12.5	600	150

* Calculated values.
C total includes input "hot" or operating capacity, socket capacity and output capacity.

A tube with a transconductance of 20 ma per volt or greater is, of course, sensitive to grid bias voltage which may be affected by grid-cathode contact potential changes. In addition to this variation, a multiplier tube is also subject to secondary-emission-ratio variation. In order to minimize the effect of both of these variables on output, the circuit for the dc supply shown in Fig. 11 has been found valuable. In this circuit,

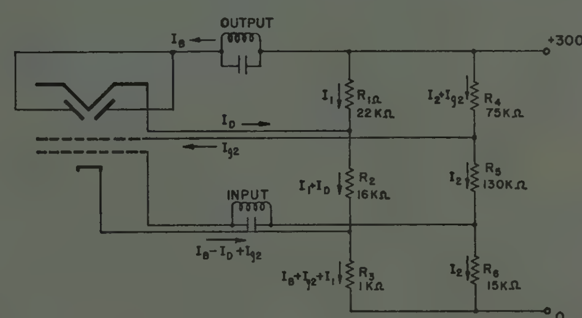

Fig. 11—Stabilizing circuit for multiplier-type tubes with typical resistance values for the tube of Fig. 7.

stabilization is obtained by running both plate and cathode currents through the self-biasing 1,000-ohm resistor, R_3. Thus variations of dynode current which change plate current will also change the bias voltage tending to keep plate current constant. Manual adjustment of control grid bias is accomplished by varying R_6. Many tests have been made using only a 100-ohm cathode resistor and no dynode current stabilization, but for maximum stability during life, better results are obtained by a circuit as shown in Fig. 11, where some typical resistance values are given.

In designing the constants of the circuit of Fig. 11, one is not completely free in the choice of R_1 and R_2. Since the current to the dynode is negative, the dynode dc voltage will always be higher than the source voltage when any series resistance is present. The requirements on R_1 and R_2 are determined from a load line on the dynode characteristic curve of Fig. 9, established so that only one stable operating point exists, i.e., the load line cuts the applicable curve at only one point. This

[9] G. T. Ford, "The 404A, a broadband amplifier tube," *Bell. Lab. Rec.*, vol. 27, pp. 59–61, February, 1949.

requirement with a margin of safety in the described tube is $R_1R_2/R_1+R_2=10{,}000$ ohms.

By using the dynode voltage source to furnish screen and anode power to other tubes in a circuit, the power lost in the potentiometer can be decreased and over-all efficiency improved. Power saving is also possible by a complete redesign of the stabilizing circuit to use a nonlinear resistance (such as Thyrite) for R_2 in the circuit, since such a resistor will be satisfactory at much higher resistance values than those needed with a linear resistor.

The most important factor in the secondary-emission amplifier tube is its life performance. The principal factors that influence the life are: bombarding current density, dynode temperature, and evaporation from the oxide cathode. The bombarding current density of 5 to 10 ma per square centimeter is kept low by spreading out the electron beam as much as possible, and dynode temperature is kept low by the large dynode area.

Control of cathode evaporation is an important factor in obtaining a long life. The amount of foreign material which can change or "poison" a secondary emitting surface is small. Johnson[10] gives an interesting example of a secondary emission ratio change of an oxide-cathode from 12 to 5 (1,250 volts bombarding energy) in 6 hours when exposed to the evaporation of another hot oxide cathode while no bombardment takes place. Johnson calculates that during this time, material equivalent to less than 0.1 monatomic layer of barium is evaporated on the cathode acting as the secondary emitting target.

In an indirectly heated oxide cathode tube, methods of evaporation control which depend upon limiting the solid angle of evaporation are not convenient, consequently the amount of evaporation from the cathode must be reduced. Two methods of decreasing evaporation from an oxide cathode are available: (1) by temperature, and (2) by chemical means.

The latter method has been principally used but, of course, any evaporation rate is quite sensitive to temperature. As is well known, in an oxide cathode, the amount of free barium produced by the reduction of the BaO coating is dependent upon the composition of the cathode core material.[11,12] The rate of evaporation of the free barium is higher by several orders of magnitude than the evaporation of barium oxide, strontium, or calcium oxide. Consequently, the life of the secondary emitter surface is a function of the amount of reducing material in the cathode core material.

In order to check the effect of reducing material in the cathode core material in the tube of Fig. 7, two lots of tubes were made and processed as nearly identically as possible. One lot had approximately 0.35 per cent reducing material in the core metal and the other had a maximum of 0.05 per cent. Both cathodes were sprayed with the same triple carbonate spray. After 500 hours of life, the secondary emission gain of the tubes with the 0.35 per cent reducing material had decayed on an average of 25 per cent, while the other lot had decayed only 6 per cent.

In any diffusion and evaporation process, temperature is an important factor. Fig. 12 shows results of temperature variation tests. A single curve which is a composite of several tubes is used to illustrate the behavior. These tests were conducted at a current density of 1 ma per square centimeter. The gain is practically constant to 2,000 hours, at which time the heater voltage of the cathode was increased to 7.5 volts, which corresponds to an increase of about 75°C in temperature. The gain immediately began to fall and continued to drop for 200 hours. On returning the heater voltage again to 6.3 volts, there is a slight increase and then a further decrease, but behavior of individual tubes varied considerably in this region. At present a tentative limit of ± 10 per cent on cathode heater voltage has been set, but it is hoped to extend this range by further work.

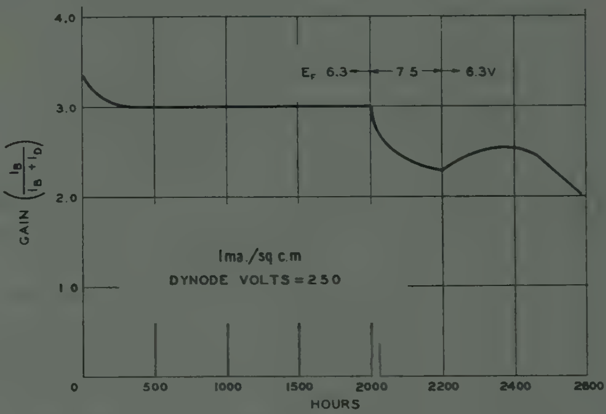

Fig. 12—Life curve illustrating the effect of cathode temperature on gain.

Conclusion

In conclusion, this work has demonstrated the successful use of secondary emitting surfaces directly exposed to oxide cathode surfaces which has greatly simplified tube construction. Possible constructions for both filament and indirectly heated cathode tubes have been successfully made and a reasonable life demonstrated.

Acknowledgment

In the course of this work, contributions were made at various times by T. Wallmark, Miss M. Mihelyi, K. R. DeRemer, M. W. Green, and M. G. Topke. The ready co-operation of K. M. McLaughlin and E. V. Space of the RCA Victor Division at Harrison, N. J., where many of the tubes were fabricated, is greatly appreciated.

[10] J. B. Johnson, "Secondary electron emission from targets of barium-strontium oxide," *Phys. Rev.*, vol. 73, pp. 1058–1073; 1948.
[11] M. Benjamin, "The influence of impurities in the core-metal on the thermionic emission from oxide-coated nickel," *Phil. Mag.*, vol. 20, pp. 1–24; 1935.
[12] E. M. Wise, "Nickel in the radio industry," Proc. I.R.E, vol. 25, pp. 714–752; June, 1937.

Propagation of Short Radio Waves Over Desert Terrain*

J. P. DAY†, MEMBER, IRE, AND L. G. TROLESE†, SENIOR MEMBER, IRE

Summary—**Results are given of an experimental investigation of the effect of relatively simple topography and meteorology upon the propagation of short radio waves over an optical 26.7-mile path and a nonoptical 46.3-mile path. Two types of meteorological conditions were encountered during the course of the experiments performed in the Arizona desert. In the daytime the atmosphere was well mixed with the index of refraction distribution nearly standard. At night a small scale duct was formed, due to a temperature inversion arising from the cooling of the ground by radiation. Measurements of the vertical distribution of field strength over a 190-foot interval were made under these two meteorological conditions for frequencies of 25, 63, 170, 520, 1,000, 3,300, 9,375 and 24,000 Mc. The effect of the diurnal meteorological cycle on the field strength is discussed for both the optical and nonoptical path. Diffraction effects on the short path due to small scale irregularities of the terrain are also discussed.**

Introduction

As THE WAVELENGTH under consideration becomes shorter, two factors add increasing complexity to the problem of the propagation of radio waves through the lower troposphere. These two factors are the inhomogeneity of the atmosphere and the irregularity of the terrain. This paper reports an experimental investigation of the influence of these two factors on the propagation of short waves varying in frequency from 25 to 24,000 Mc for conditions found over desert terrain.

The experiments described in this paper were performed during the winter season in the Arizona desert where the atmosphere is usually dry and clear with little cloudiness. Due to nocturnal radiation from the ground, a large diurnal change in surface temperature takes place. This nighttime cooling of the ground results in a temperature inversion in the lower layers of the atmosphere, thus forming a small scale radio duct. The duct increases in height and intensity as the night progresses. A series of index of refraction profiles for a 24-hour period is shown on the bottom chart of Fig. 2, Although the duct is of weak intensity and low height the effect on the higher frequencies is pronounced. In the daytime the atmosphere is well mixed and the index of refraction distribution is nearly linear with height, except for a steep positive gradient very near the surface due to intense heating of the ground. These two conditions, nearly standard refraction in the daytime and a small scale duct at night, were the only conditions encountered during the investigations.

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The terrain was remarkably uniform with irregularities which would ordinarily be considered of very small scale; however, the effect of these irregularities is found to be appreciable at the higher frequencies.

Experimental Setup and Procedure

Measurements were made over two paths, one 26.7 miles in length, and the second a longer path of 46.3 miles. A few measurements were made by means of an airplane to altitudes of 8,000 feet. The common transmitting terminal was located near Gila Bend, Ariz., the 26.7-mile path receiving station near Sentinel, Ariz., and the 46.3-mile receiving terminal near Dateland, Ariz. The two paths were nearly co-linear, with the 26.7-mile receiving station displaced 600 feet from the long path. The terrain was mostly flat desert covered with small brush. Profiles of the two paths are shown in Fig. 1. An examination of the profiles shows that the deviations of the terrain from a smooth sphere are less than 100 feet for both paths.

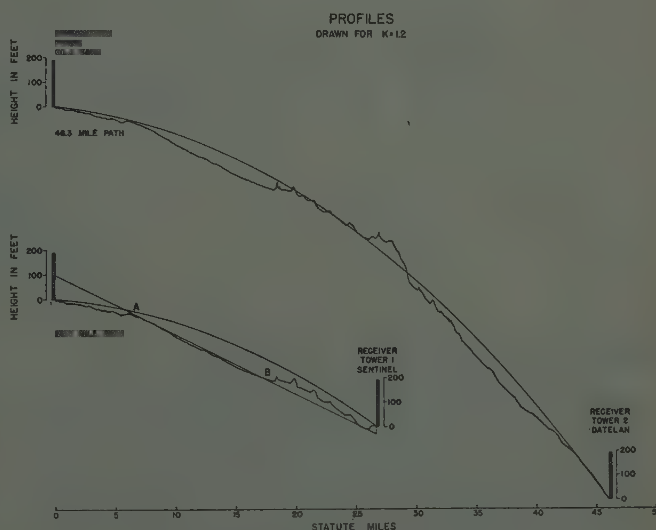

Fig. 1—Profiles of transmission paths.

Transmitters and receivers were located in elevator cabs mounted on 200-foot towers, thus enabling the measurement of vertical distribution of field strength. On the short path, receivers were within the line of sight for most transmitter and receiver elevations, while on the longer path, receivers were always well below the line of sight. The vertical distributions of field strength

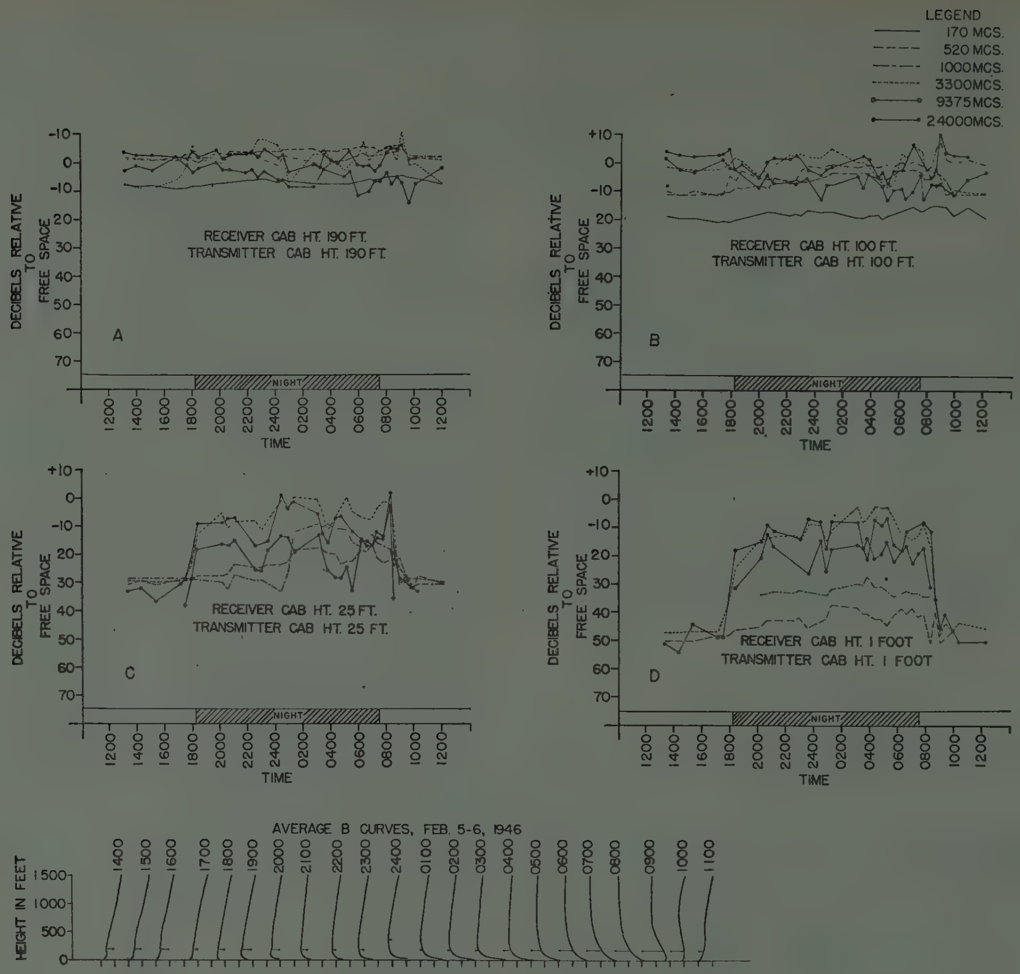

Fig. 2—Diurnal variation of field strength and index of refraction distribution for the 26.7-mile path.

Arrow Indicates value $295(B-1)\times10^6$ Units

Scale

290 300 310

$(B-1)\times10^6$

Note

$B=n+ah$

n = Index of refraction

$a=1.18\times10^{-8}$ per foot

h = Height in feet.

(height-gain curves) were recorded simultaneously for 170, 520, 1,000, 3,300, 9,375, and 24,000 Mc at the two receiver locations. Later a few additional measurements were added at 25 and 63 Mc. For 3,300, 9,375, and 24,000 Mc the transmitters were modulated with one-microsecond pulses and the received field recorded by peak-pulse reading receivers. Continuous wave was used for the lower frequencies. Height-gain curves were obtained at frequent intervals during 24-hour operating periods. Most of the measurements were made with horizontal polarization with occasional checks with vertical polarization.

Meteorological measurements were made concurrently with radio measurements. Dry-bulb temperature and dew-point temperature were measured at various heights up to 200 feet on each tower, and dry-bulb temperature and relative humidity were measured to heights of 1,500 feet at three intermediate points along the path by means of a captive balloon technique. These measurements furnished the necessary data for calculation of the index of refraction distribution.

Results of Measurements

Diurnal Variations in Field Strength

Measurements were taken during a number of twenty-four hour intervals over a period of several months. The cyclic variation in meteorological conditions occurs with the same general character night after night and, in fact, from year to year. The field-strength variations reflect this consistency. Apart from transient details it suffices, therefore, to examine representative samples for one 24-hour interval. To illustrate the diurnal variation, discrete points were taken from the height-gain curves and plotted for several fixed terminal heights. This type of plot is shown for the 26.7-

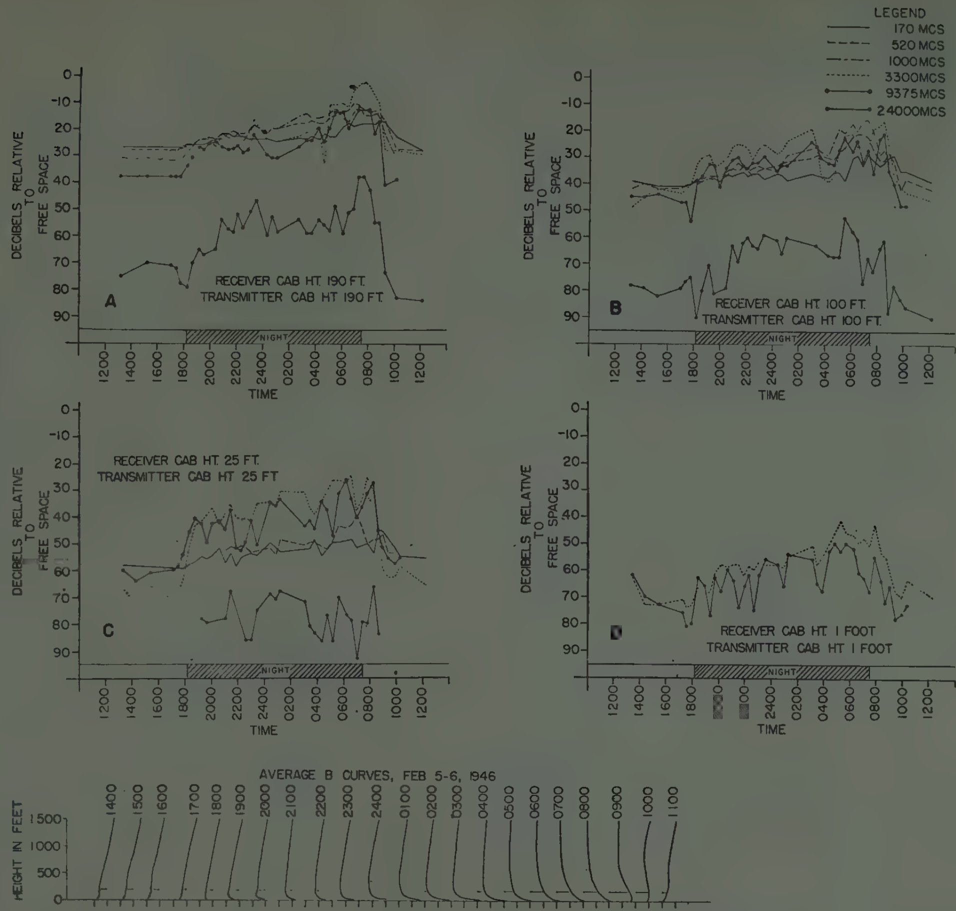

Fig. 3—Diurnal variation of field strength and index of refraction distribution for the 46.3-mile path.

Arrow Indicates Value 295(B-1)$\times 10^6$ Units

Scale

290 300 310

(B-1)$\times 10^6$

Note

$B=n+ah$

n = Index of refraction

$a=1.18\times 10^{-8}$ per foot

h = Height in feet.

mile path in Fig. 2. The terminal heights are given as heights of the elevator cab floors above the base of the tower. The antennas were located on the face of the cab from 2 to 7 feet above the floor. Six frequencies are shown, as well as an hourly plot of the index of refraction distribution. The index of refraction curves are averaged curves of equivalent index of refraction (B curves) based on the transformation of the earth's radius to 4/3 times the actual radius.[1] This multiplying factor K has the value 4/3 for certain average conditions for which the atmosphere's refraction is termed standard. During the daytime the index of refraction curves show a steep positive gradient in the first few feet, and are nearly linear above with a slight positive slope. Thus, apart from the first few feet, the atmosphere is slightly substandard during the daytime and equivalent to $K=1.2$ rather than 4/3. From the profile of Fig. 1 which is drawn for $K=1.2$, it will be seen that the 26.7-mile path is optical for the terminal heights of 190 feet, Fig. 2(a), and 100 feet, Fig. 2(b), but is nonoptical for the terminal heights of 25 feet, Fig. 2(c), and 1 foot, Fig. 2(d). An examination of Fig. 2 shows that there is a marked difference in the diurnal change of field strength for the optical terminal heights as contrasted to the nonoptical heights. This is especially true for the microwave frequencies. The higher terminal heights are in the interference region and the fields may decrease or increase as the temperature inversion forms, depending on how the lobe structure is distorted for the various frequencies. For the low terminal heights the field always increases as the duct

[1] $B=n+\alpha h$ where n is the actual index of refraction, h is the altitude in feet, and $\alpha=1.18\times 10^{-8}$ per foot.

forms, and, though variable during the night, is always higher at night than during the day. The magnitude of the change is always much greater for the non-optical heights.

Receivers on the 46.3-mile path are well below the line of sight for all terminal heights. Fig. 3 shows the character of the diurnal change of field strength for this path as well as the index of refraction curves for a 24-hour period. An examination of the index of refraction curves of Fig. 3 shows that the bottom few feet of the index of refraction profile swings from a steep positive slope during the hot daytime hours to a negative slope between 1700 and 1800 somewhat before sunset which occurred at 1810. Above the first few feet the curve varies linearly with height, and has a slight positive slope during the afternoon and early evening. As the night progresses from 1800 to 0900 the following morning, the duct resulting from the low-level negative gradients is constantly rising in height and increasing in intensity. During this period the upper portion of the meteorological profile is gradually changing from a positive slope toward a negative slope. From 0800 to 0900 in the period after sunrise (at 0725), the ground begins to heat and the slope in the bottom few feet changes back to positive. Above the first few feet the slope is negative for several hundred feet, but gradually swings over as the morning progresses to form finally a nearly linear curve of slightly positive slope which persists throughout the afternoon.

The diurnal meteorological cycle is more readily apparent in the top plot of Fig. 4. This shows contours of constant index of refraction gradient plotted against altitude and time. Positive gradients are shown by dotted lines and negative gradients by solid lines. At low altitudes the positive gradients die out after 1700 and negative gradients appear and persist throughout the night. The field strength plots of Fig. 4 are plots of the maximum values of field recorded in the 190-foot vertical receiver excursion for a transmitter cab height of 100 feet on the 46.3-mile path. This plot shows that the rise and fall of the signals correlate with the onset and disappearance of negative gradients at low altitudes. Some of the minor signal variations can be associated in time with corresponding variations in the gradient structure, but the correlation is not consistently one to one.

On the 46.3-mile path the diurnal change in field strength increases as the frequency increases to 3,300 Mc but it is less for 9,375 and 24,000 Mc than for 3,300 Mc.[2]

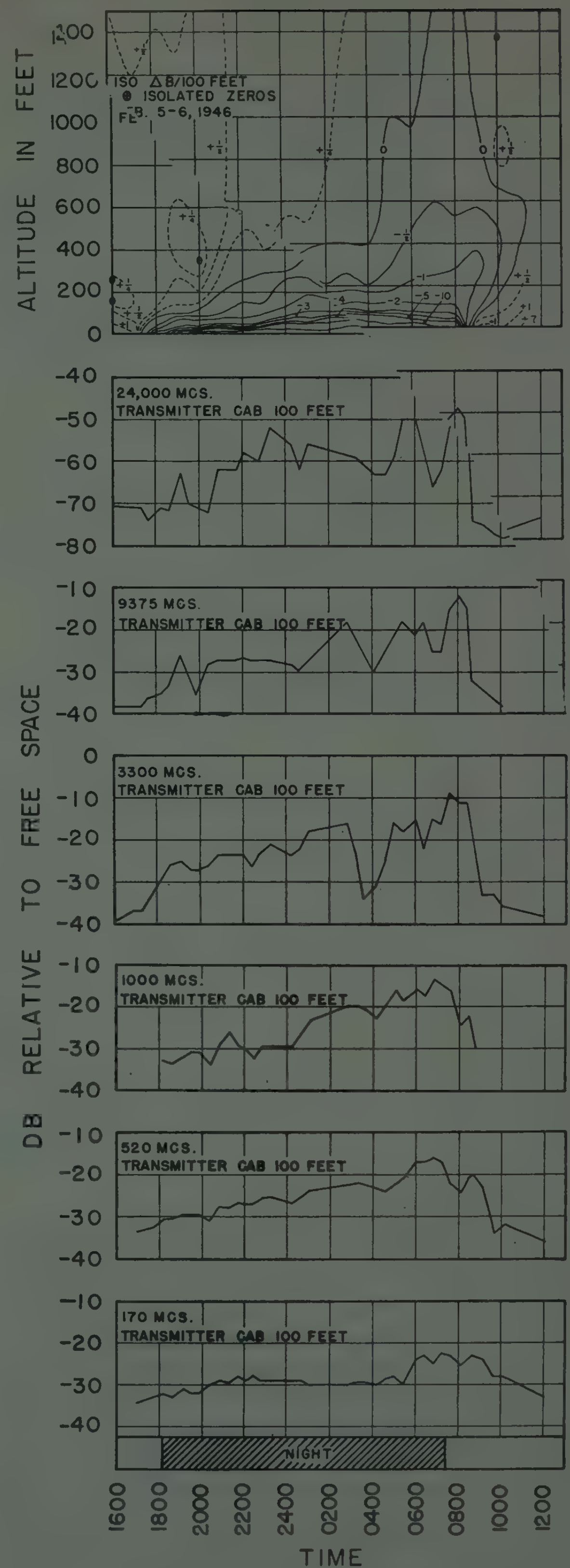

Fig. 4—Diurnal variation in field strength and index of refraction gradient.

This result is unexpected as is also the fact that the diurnal change for 3,300 and 9,375 Mc is greater on the

[2] Relative values of field strength were measured with good accuracy on all frequencies. Absolute values of field are considered good except for 24,000 Mc. The meteorological condition is most nearly the same from day to day at a time in the middle of the afternoon. On the 26.7-mile path, at terminal heights for maximum field, the measured field at this time of day varied less than ±2.5 db from a mean value over a two-month period for all frequencies except 24,000 Mc. On 24,000 Mc the observed variation was ±13 db. What part of these variations is due to atmospheric effects and what part to measurement inaccuracy is not known, but these figures are considered upper limits of experimental error.

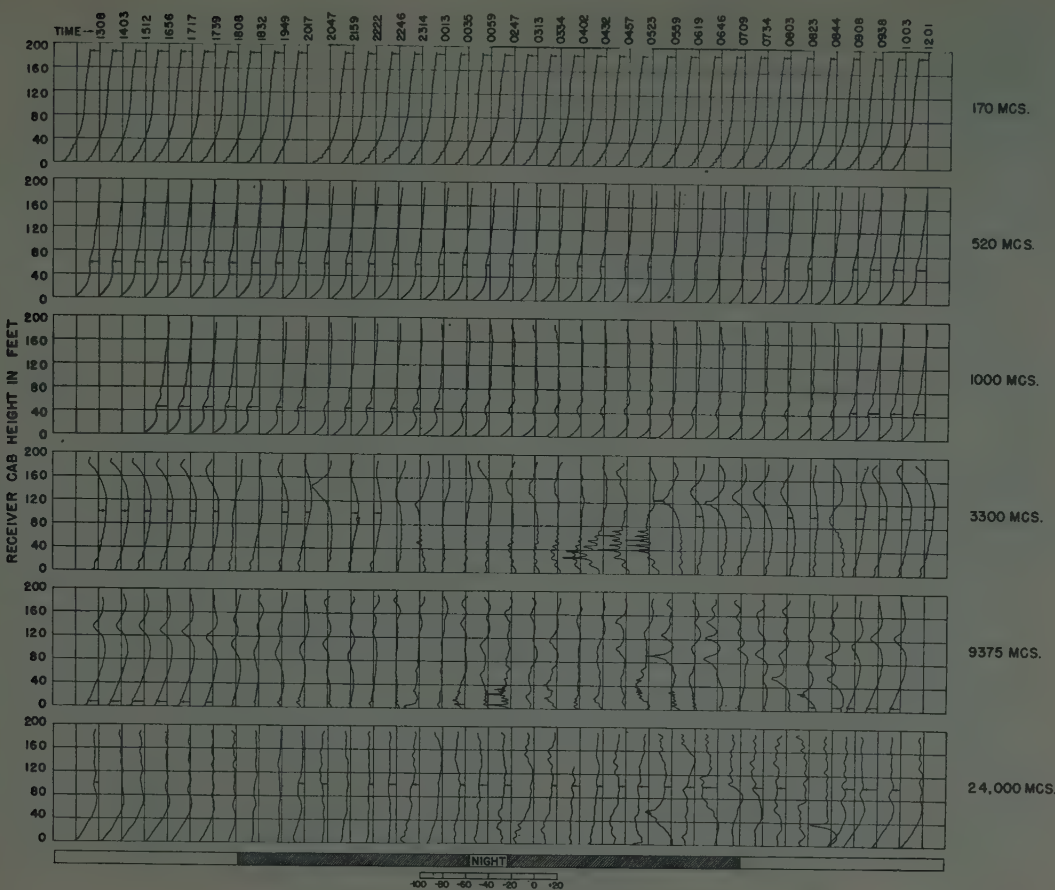

Fig. 5—Twenty-four hour cycle of received height-gain curves for the 26.7-mile path with transmitter cab at 190 feet. The time at which each curve was taken is indicated above the zero-db reference line for field strength. An arrow is drawn from each zero-db reference line to the curve for which it applies. Field-strength scale in db, relative to free-space field, is shown at the bottom of the figure.

26.7-mile path, with low terminals, than on the longer path. (The 24,000-Mc signal was below detection at low heights and cannot be included in this comparison.) Rocco and Smyth in a paper[8] on the subject of diffracted fields have indicated a possible explanation of the above facts. During the daytime, when the atmosphere is nearly standard, the received fields on the 46.3-mile path should agree with values calculated on the basis of standard diffraction theory. Agreement was found on the lower frequencies, but for low terminal heights the measured signals at 3,300 and 9,375 Mc were many tens of db too high. Apparently some mechanism other than refraction and diffraction must be postulated to explain these high daytime fields, and atmospheric scattering by turbulent air parcels has been suggested as a possible mechanism. Meteorological measurements have been made on these turbulent air parcels, but the significant connections between the radio and meteorological observations of this phenomenon are as yet lacking.

During the night, at times of greatest refraction, the fields for 3,300 and 9,375 Mc on the long path rise to about the same absolute value. In the daytime the observed fields for 9,375 Mc are in general somewhat higher than for 3,300 Mc (for low terminals). The calculated diffracted field on the other hand is lower for 9,375 Mc than for 3,300 Mc. The minimum value of field strength in a 24-hour cycle drops to the daytime diffracted field value for lower frequencies, but for high frequencies the "scattered" field is dominant and prevents the signal from dropping to the diffracted field value. The scattering mechanism is apparently more effective on 9,375 Mc than on 3,300 Mc, and consequently the diurnal change on 9,375 is less. Table I illustrates

TABLE I

CALCULATED AND OBSERVED FIELD-STRENGTH VALUES IN db RELATIVE TO FREE SPACE

46.3-Mile Path	*Transmitter and Receiver Heights*		*25 Feet*
Frequency in Mc	*(1) Calculated Diffraction Field*	*(2) Observed Daytime Field*	*(3) Maximum Observed Nighttime Field*
3,300	−94	−70	−25
9,375	−127	−60	−27

[8] M. D. Rocco and J. B. Smyth, "Diffraction of high-frequency radio waves around the earth," PROC. I.R.E., vol. 37, pp. 1195–1204; October, 1949.

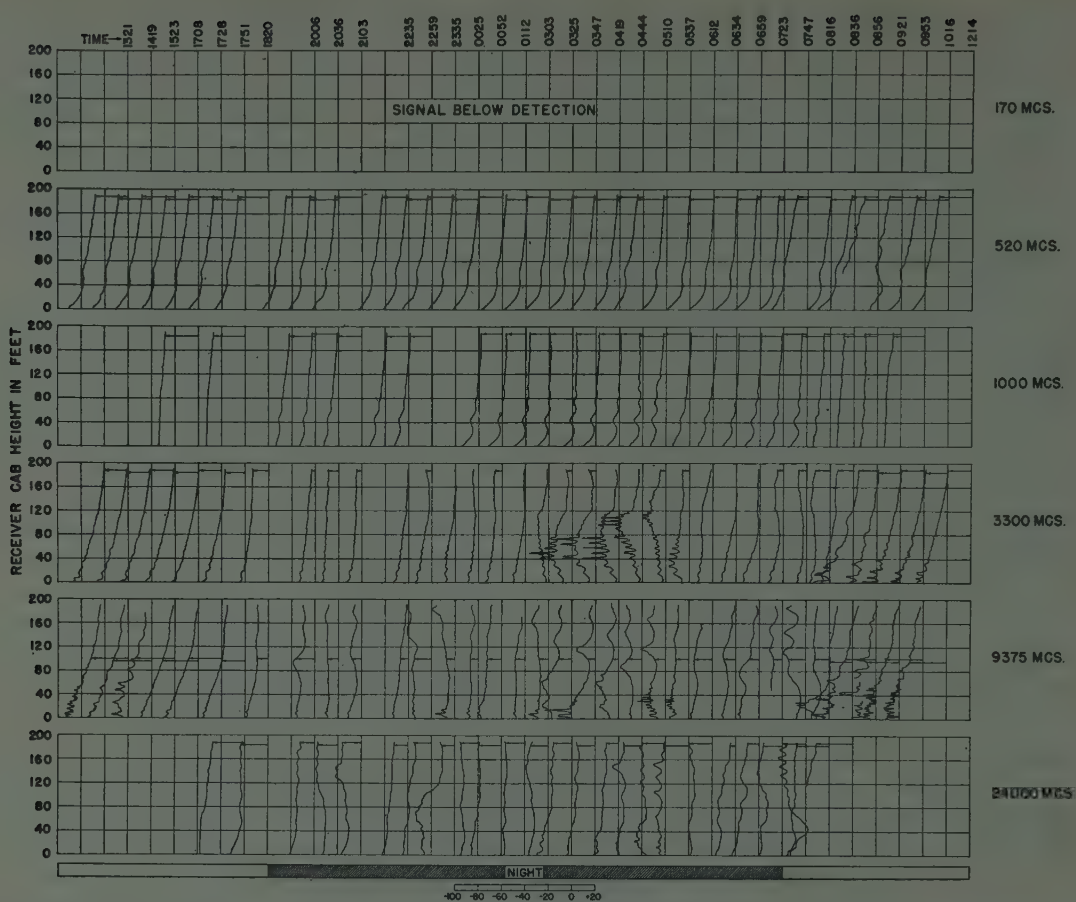

Fig. 6—A twenty-four hour cycle of received height-gain curves for the 26.7-mile path with transmitter cab at 1 foot. The time at which each curve was taken is indicated above the zero-db reference line for field strength. An arrow is drawn from each zero-db reference line to the curve for which it applies. Field-strength scale in db, relative to free-spaced fiel, is shown at the bottom of the figure.

this point with representative values of field strength for transmitter and receiver heights of 25 feet.

The fact that the observed diurnal change is higher on the 26.7-mile path with low terminals than on the 46.3-mile path implies that the scattered field component is much weaker than the diffracted field on the short path.

Vertical Distribution of Field Strength

26.7-Mile Path

The vertical distribution of field strength is shown as height-gain curves for one 24-hour period on the 26.7-mile path by Figs. 5 and 6, which are for transmitter cab heights of 190 and 1 foot, respectively. Height-gain curves for six frequencies were recorded simultaneously.

During the daytime the shape of the height-gain curves is in general repetitive for all frequencies, although for the low transmitter height the higher frequency curves show minor variations. With the transmitter cab at 190 feet (Fig. 5), the first interference maximum for the three lower frequencies occurs at receiver heights considerably above the 190-foot tower height. At 3,300 Mc the received daytime height-gain curve shows the first interference maximum at about 110 feet and the first minimum just below 190 feet. For 9,375 Mc two interference maxima and one minimum are noted. The 24,000-Mc daytime height-gain curve shows but slight evidence of an interference pattern. It will be noted that other small scale maxima and minima are present at all frequencies. For example, the 1,000-Mc curve has maxima at 30 feet, 60 feet, 85 feet, and minima at 45 feet, 75 feet, and 100 feet. These variations are characteristic of a knife-edge diffraction pattern and will be discussed in detail later in a section on diffraction and interference effects.

Just before sunset the shape of the height-gain curves for the three higher frequencies begins to be modified. For a transmitter cab height of 190 feet a gradual depression of the lobe structure takes place on 3,300 and 9,375 Mc, but early in the evening any semblance of a lobe structure disappears and the shape of the curves is then quite variable throughout the night. At times the field is almost constant with height. In other cases a deep minima may be present on one curve and absent on the next. Some time after sunrise the curves for 3,300 and 9,375 Mc again show the daytime interference lobe structure with lobes tending to rise as the morning progresses.

By contrast with the higher frequencies, the height-gain curves for 170, 520, and 1,000 Mc are only slightly

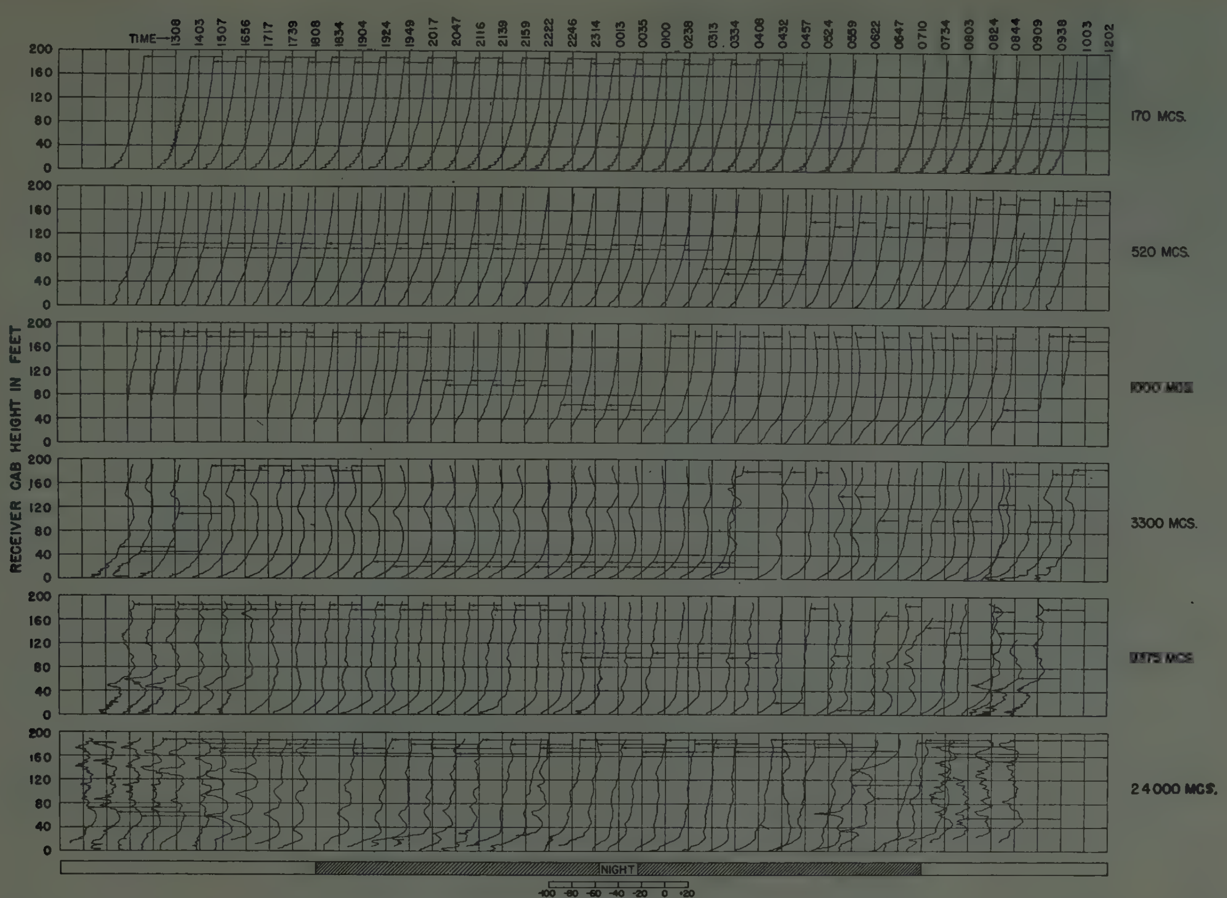

Fig. 7—A twenty-four hour cycle of received height-gain curves for the 46.3-mile path with transmitter cab at 190 feet. The time at which each curve was taken is indicated above the zero-db reference line for field strength. An arrow is drawn from each zero-db reference line to the curve for which it applies. The field-strength scale in db, relative to the free-space field, is shown at the bottom of the figure.

modified by the presence of the nighttime duct. The lower portions of the curves gradually shift toward higher field strength as the night progresses, while the upper portions are but little affected.

46.3-Mile Path

The height-gain curves for a 24-hour period on the 46.3-mile path are presented in Figs. 7 and 8. On this longer path the shape of the daytime height-gain curves is more variable than on the short path, although the general shape is repetitive even on the higher frequencies. The curves for the lower frequencies are altered somewhat more during the night on the long path than on the short path. The curves for 3,300 and 9,375 Mc, on the other hand, change less throughout most of the night on the long path than on the short one.

Receivers on the 46.3-mile path are well below the line of sight and the height-gain curves should show the field increasing monotonically with height, especially in the daytime when the atmosphere is nearly standard. However, from Fig. 7 it is apparent that the curves for 3,300, 9,375, and 24,000 Mc show pronounced minima both in the daytime and nighttime. On 3,300 Mc a minimum occurs at a height which is about 130 feet in the daytime, and then drops to about 120 feet and persists for a good part of the night. The height of this minimum seems to be insensitive to transmitter height. This suggests that the minimum is due to some type of diffraction process near the receiver location, but the specific nature of the process is not known.

At low transmitter height, Fig. 8, it will be noted that the height-gain curves for 3,300 and 9,375 Mc show rapid scintillations of field in the daytime. These are variations with time, not with height, and appear when receivers are fixed in height or moving. At the low transmitter height, the average field is at times practically constant with height. It is the presence of these scintillating fields, which are much higher than predicted by standard diffraction theory, which suggests a scattering process in the atmosphere.

Diffraction and Interference Effects

In this section the effect on the lobe structure of the small terrain irregularities of the short path will be considered. An examination of the profile of the 26.7-mile path shows irregularities in the terrain which,

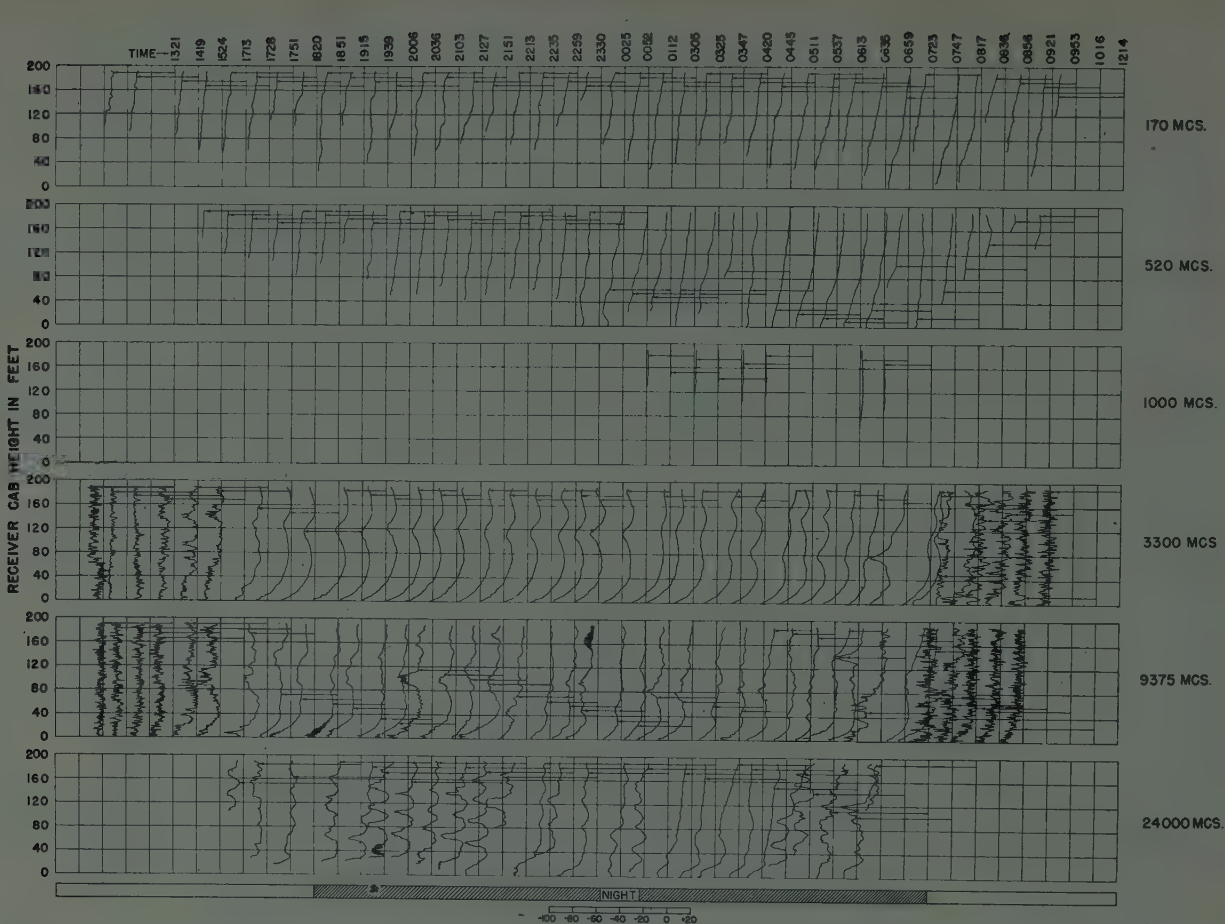

Fig. 8—A twenty-four hour cycle of received height-gain curves for the 46.3-mile path with transmitter cab at 1 foot. The time at which each curve was taken is indicated above the zero-db reference line for field strength. An arrow is drawn from each zero-db reference to the curve for which it applies. The field-strength scale in db, relative to the free-space field, is shown at the bottom of the figure.

though small, might produce diffraction effects somewhat like knife-edge diffraction.[4,5] The field-strength profiles of Fig. 5 show variations with height which have the appearance of a knife-edge diffraction pattern. These variations are persistently present, day and night, and for all transmitter heights. An examination of the heights at which minima and maxima occur seems to confirm that the behavior is like knife-edge diffraction.

In Fig. 9 the experimental data are shown in the form of a plot of wavelength versus the square of the heights at which the maxima and minima occur. For fixed wavelength the height of successive maxima and minima can be shown as a function of the index integer ($n=1$ for first maximum, $n=2$ for first minimum, etc.). Fig. 10 shows this type of plot for 3,300 Mc where many maxima and minima are noted. Both of these plots[6]

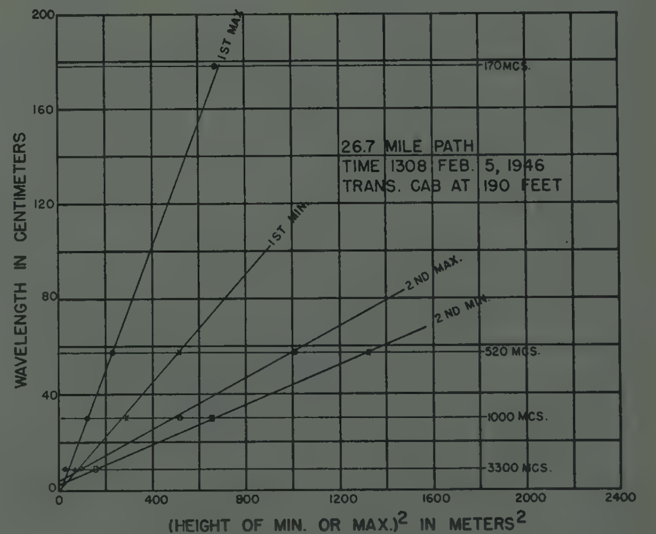

Fig. 9—The dependence of height of diffraction maxima or minima on wavelength.

show that the data conform with Fresnel-Kirchoff diffraction theory for a knife edge.

The location of the apparent knife edge was calculated from the diffraction pattern at the different fre-

[4] J. C. Shelleng, C. R. Burrows, and E. B. Ferrell, "Ultra-short-wave propagation," PROC. I.R.E., vol. 21, pp. 427–463; March, 1933.

[5] Summary Technical Report of the Committee on Propagation, N.D.R.C., vol. 1, chap. 14; vol. 3, chap. 8, 1946.

[6] In both Fig. 9 and Fig. 10 a better fit was obtained by adding a constant factor of 12 feet to the cab height. Since the base of the tower is a few feet above the general level of surrounding terrain, and antennas are a few feet above the cab floor, this seems like a reasonable correction to the arbitrarily chosen height scale.

quencies. Values ranging from 0.4 to 0.7 mile from the receiver tower were obtained. An examination of the profile shows a rolling ledge at 0.5 mile from the receiver tower (26.2 miles from the transmitter tower) rather than the expected sharp ridge. Measurements taken later with a 3,300-Mc transmitter located in a truck at several distances confirmed that the ledge was responsible for the diffraction pattern. Fig. 11 shows successive height-gain curves taken at 0.32, 0.68, 2.0, and 6.9 miles from the receiver tower. Curve *A* at 0.32 mile shows the typcal interference pattern with equispaced lobes due to a direct and ground-reflected ray. In contrast, curves *B*, *C*, and *D* taken with the transmitter located beyond the ledge show striking diffraction patterns with the heights of successive maxima and minima varying as the square root of the index integer.

As mentioned before, this diffraction effect was persistently present and generally did not vary with time of day. Occasionally, however, a striking intensification of the diffraction pattern was noted during the night. One example can be seen on Figs. 5 and 6 for the 3,300 and 9,375-Mc frequencies at times between 0330 and 0530. On Fig. 10 the heights of successive maxima and minima versus the index integer are shown as crosses for daytime data, dots for nighttime values. The only difference between the night and day data is a slight shift in the heights. The variations seem unquestionably to be due to the same diffraction effect, but with tremendous increase in the amplitude of the variations (up to 23 db). The cause of this effect is not understood.

The daytime height-gain curves for 24,000 Mc with transmitter cabs at 190 feet have a shape suggestive of a second knife-edge diffraction pattern with the knife edge located at greater distance from the receiver tower than the ledge at 0.5 mile. The profile shows two sharp ridges at 6.8 and 8.3 miles from the receiver tower (19.9 and 18.4 miles from the transmitter). To the east of these ridges (left on profile plot) the terrain is nearly a plane for about 12 miles between points *A* and *B*.

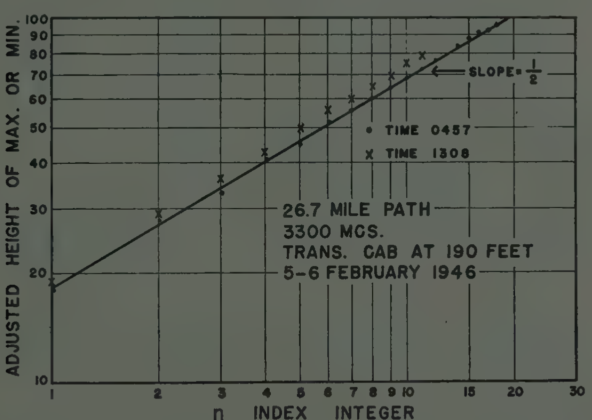

Fig. 10—A comparison of night and day diffraction pattern showing dependence of height of maxima and minima on the index integer.

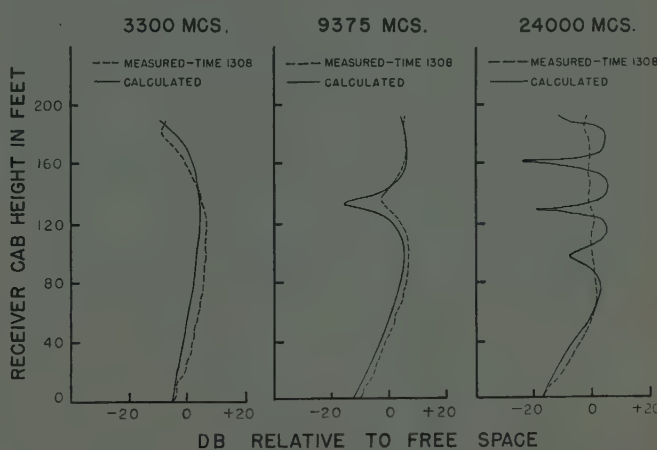

Fig. 12—A comparison of calculated and measured height-gain curves for a 26.7-mile path, with transmitter cab at 190 feet. Unity reflection coefficient was used in the calculations.

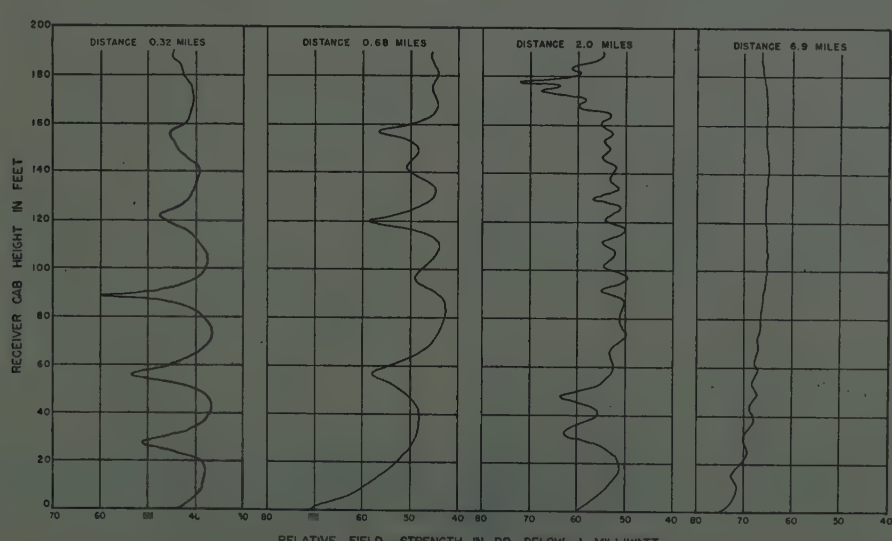

Fig. 11—Height-gain curves taken with mobile transmitter on either side of a diffracting ledge.

An extension of this plane intersects the transmitter tower at 97 feet and the receiver tower 48 feet below its base. Height-gain curves were calculated for transmitter heights of 190 feet assuming a direct ray and a reflected ray each ray being perturbed in amplitude and phase by a single knife edge located at 19.9 miles from the transmitter. If a reflection coefficient of unity is assumed at the ground the calculated curves depicted in Fig. 12 are obtained for the three microwave frequencies. The agreement with the measured curve is fair for 3,300 Mc. For 9,375 and 24,000 Mc the interference minima are much too deep as compared to measured values. It is interesting to note, however, that the minima on 9,375 and 24,000 Mc are filled in to some extent, even though a reflection coefficient of unity was used in the calculations. Thus if the height-gain curve was assumed to result from two rays unperturbed by knife-edge diffraction, the apparent reflection coefficient estimated from the height-gain curve would be less than the actual reflection coefficient. For example, the calculated curve for 24,000 Mc would indicate an apparent reflection coefficient increasing with receiver height, even though unity reflection coefficient was used in the calculations. If diffraction is not taken into account and the observed height-gain is assumed to be made up of a direct and ground reflecting ray, the observed variation of the depth of minimum with height might be interpreted as a variation of the reflection coefficient.[7]

A reflecting surface may be considered flat if surface irregularities do not cause path differences exceeding a small fraction of a wavelength. Thus specular reflection may be expected if

$$\frac{H \cos \theta}{\lambda} \ll 1 \tag{1}$$

where H is the height of surface irregularities, λ the wavelength, and θ the angle of incidence. The measurements of Ford and Oliver[8] indicate that when the above fraction is about $\frac{1}{8}$ a reflection coefficient of 0.5 results, and that when the fraction is $\frac{1}{2}$ a reflection coefficient greater than 0.1 is improbable. On this basis reflection coefficients of 0.6 for 9,375 Mc and 0.2 for 24,000 Mc appear to be plausible values if irregularities of 10 feet are assumed in the reflecting plane. Between points A and B on the profile, where reflection takes place, the terrain is plane within ± 10 feet. Using these values of reflection coefficient, the curves shown in Fig. 13 were calculated which are in fair agreement with measured curves.

For the lower frequencies, even the irregularity at 19.9 miles which was considered as a knife edge for the higher frequencies satisfies inequality (1). In this case calculations based on a smooth earth which best fits the profile,

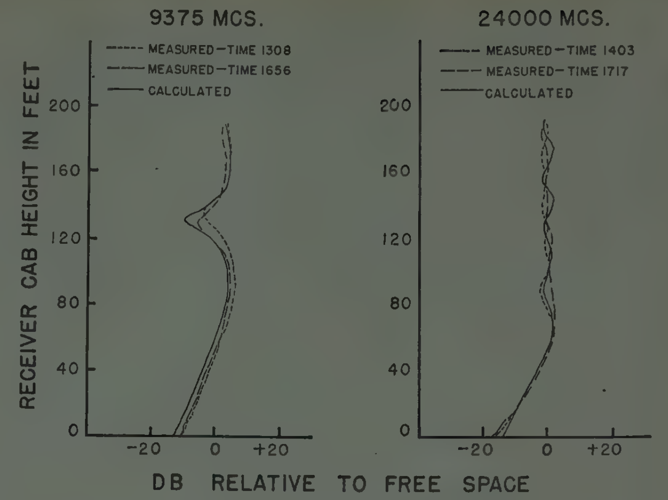

Fig. 13—A comparison of calculated and measured height-gain curves for a 26.7-mile path, with a transmitter cab at 190 feet. A reflection coefficient of 0.6 was used for 9,375 Mc and 0.2 for 24,000 Mc

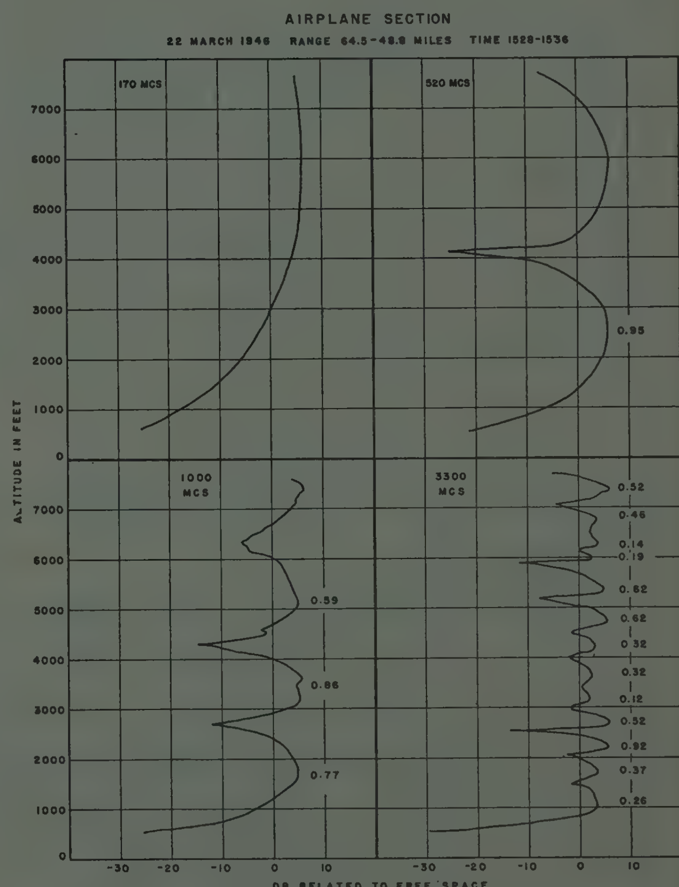

Fig. 14—Height-gain curves taken with receivers in an airplane. The transmitter cab was at 190 feet.

and a high value of reflection coefficient would seem reasonable. This is borne out by Fig. 14, which shows height-gain curves taken by means of an airplane. Field strength was recorded as the plane descended from an altitude of 7,700 feet to 600 feet. Although the angle

[7] E. W. Hamlin and W. E. Gordon, "Comparison of calculated and measured phase difference at 3.2 centimeters wavelength," Proc. I.R.E., vol. 36, p. 1218; October, 1948.

[8] L. H. Ford and R. Oliver, "An experimental investigation of the reflection and absorption of radiation of 9-cm. wavelength," *Proc. Phys. Soc.* (London), vol. 58, pp. 265–280; May, 1946.

of incidence is much smaller for this airplane data as compared to the tower data on the 26.7-mile path, the height-gain curves for 170 and 520 Mc are very smooth with high apparent reflection coefficient, indicating insensitivity to irregularities of the ground. At 1,000 Mc the height-gain curve appears to be affected somewhat by ground irregularity, and at 3,300 Mc the curve is quite irregular with the apparent reflection coefficient varying from 0.12 to 0.92, depending upon the degree of roughness of ground at the point where reflection takes place.

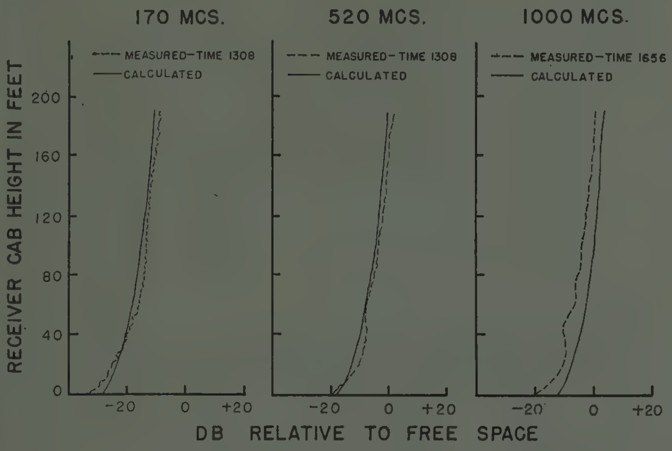

Fig. 15—A comparison of calculated and measured height-gain curves for a 26.7-mile path, with transmitter cab at 190 feet. The calculation was made on a basis of plane earth and unity reflection coefficient.

Calculated and measured height-gain curves for the three lower frequencies are shown in Fig. 15 for the 26.7-mile path. For these calculations a plane earth was assumed, since deviations from a plane along the path still satisfy the inequality (1) for 170 and 520 Mc at the angles of incidence involved. The agreement for these two frequencies, apart from the effects of the knife edge at 0.5 mile from the receivers, is fairly good. For 1,000 Mc the agreement is not as good; the shape of the curve is about right, but the absolute value is displaced. This may be due partly to possible inaccuracy of the measurement of absolute value. However, it is apparent that for this frequency the irregularities of terrain on this path may begin to have an effect.

For transmitter heights less than 190 feet other irregularities in the terrain would probably have to be considered. No attempt was made to calculate height-gain curves for transmitter heights other than 190 feet.

Conclusions

Some of the main features brought out by these experiments may be summarized as follows:

In the Arizona Desert during the winter season, which is characterized by clear skies and low moisture content in the atmosphere, nocturnal radiation from the ground can produce a duct which has a marked effect on the propagation of short radio waves. The scale of the meteorological phenomena is such that the diurnal change of radio fields varies from a negligible value on 63 Mc to about fifty decibels at microwave frequencies on paths of 26.7 and 46.3 miles. Thus this type of meteorological situation primarily affects microwave frequencies in contrast to larger scale meteorological conditions found in other areas, particularly coastal areas which have a pronounced influence on the propagation of frequencies as low as 50 Mc.

On the 26.7-mile path the diurnal variation of signals is markedly different for high terminals, within the line of sight, than for low terminals below the line of sight. For high terminals the fields vary less, and at the higher frequencies may decrease rather than rise at night. For low terminals the signals rise sharply at night for all frequencies, and display a much greater diurnal change.

For the 46.3-mile path the maximum diurnal change occurs for 3,300 Mc with less change at 9,375 and 24,000 Mc. This unexpected result arises because, in the daytime when conditions are nearly standard, the fields for 9,375 and 24,000 Mc do not drop to values which would be expected on the basis of diffraction around a suitably enlarged earth. It seems plausible that some mechanism other than diffraction, such as scattering from inhomogenous air parcels, plays a role at these higher frequencies for long paths. Such inhomogeneities apparently do not influence the field strength on the short optical path.

A distinct correspondence is noted between the formation and breaking up of the temperature inversion and the diurnal variation of field strengths. However, a detailed one to one correlation is not apparent.

For the 170, 520 and 1,000 Mc frequencies the effect of the duct on the shape of the height-gain curves is not pronounced. The major effect is a shift in absolute value. In contrast, the microwave height-gain curves are radically modified in shape during the night.

Although the terrain in this particular location is exceptionally smooth and regular, rather striking effects due to small irregularrities were noted. In one case on the 26.7-mile path a small ledge, very unlike a knife edge in appearance, is shown to produce a diffraction pattern which conforms fairly well to simple knife-edge theory. Occasionally at night an extreme intensification of this diffraction pattern occurred. No explanation has been found for this apparent change of magnitude of a diffraction effect by the presence of a refractive atmosphere.

On the optical 26.7-mile path the irregularities of the terrain appreciably perturb the microwave fields, whereas the lower frequencies are fairly insensitive to the small irregularities found on this path.

Spurious Modes in Coaxial Transmission Line Filters*

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Summary—**Coaxial transmission line filters in which the shunt reactive elements are conducting rods inserted between the inner and outer conductors of the series coaxial transmission line may be analyzed by the use of well-known transmission line equations if only the *TEM* mode propagates. When other "spurious" modes are important, the problem is not so simple. Some equations are presented for the cutoff frequencies of such spurious modes, and supporting experimental data is included.**

I. Introduction

COAXIAL TRANSMISSION line filters for use at very high frequencies have been described[1-3] and their elementary theory developed. It has been shown[1] that for very high cutoff frequencies the filter shunt transmission lines become extremely short. If these shunt transmission lines are omitted, the elementary low-pass and band-pass filter prototypes become those shown in Figs. 1 and 2, respectively. Although the shunt elements indicated are single rods, it is often of advantage to use several rods which may be grouped in various ways around the circumference of the filter body.

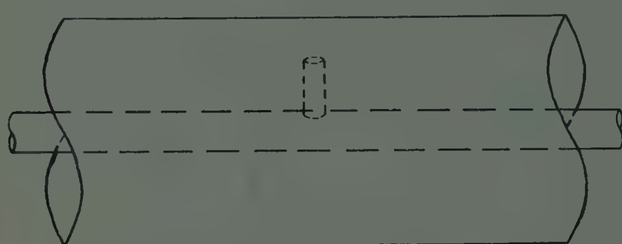

Fig. 1—Low-pass filter prototype with shunt transmission lines omitted. (Midseries section).

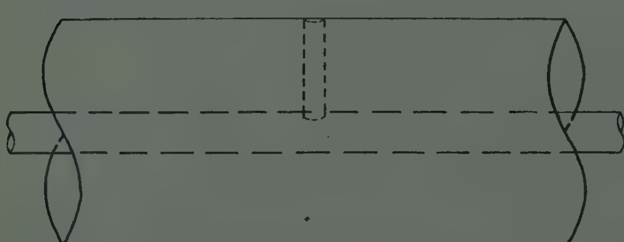

Fig. 2—Band-pass filter prototype with shunt transmission lines omitted. (Midseries section).

II. Band-Pass Coaxial Transmission Line Filters

The simple band-pass filter of Fig. 2 has an electrical equivalent circuit shown in Fig. 3, where the series transmission line has been replaced by its exact equivalent "pi" and the shunt rod by a transmission line representation. If more than one shunt rod is used per section, Fig. 3 is still applicable if the shunt rod group is translated into an equivalent single transmission line. The reactive elements are defined as follows:

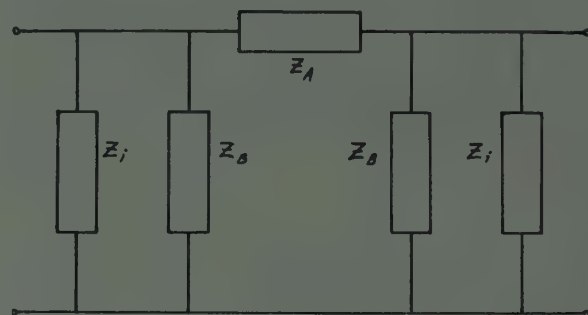

Fig. 3—Equivalent circuit of the band-pass filter prototype shown in Fig. 2. (Midshunt section).

$$Z_A = jZ_k \sin 2\phi f/f_k \tag{1}$$

$$Z_B = -jZ_k \cot \phi f/f_k \tag{2}$$

$$Z_i = jZ_{01} \tan \theta_1 f/f_k \tag{3}$$

where Z_k and Z_{01} are the characteristic impedances of the series and equivalent shunt transmission lines respectively, θ_1 is the electrical length of a shunt rod at the frequency f_k, 2ϕ is the electrical length of the series line at the frequency f_k, and f_k is an arbitrary reference frequency.

The lower cutoff frequency of the band-pass filter occurs when the ratio $Z_1/4Z_2$ (Z_1 and Z_2 are the total series and shunt impedances of a filter section, respectively) assumes its lowest frequency zero. Making the substitution,

$$Z_{01} = Z_k/d \tag{4}$$

the lower cutoff frequency is given by the smallest value of f satisfying the equation

$$d \cot \theta_1 f/f_k = \tan \phi f/f_k. \tag{5}$$

The characteristic impedance Z_{01} is calculated by an empirical formula given previously[1] for low-pass filters. This is,

$$Z_{01} = 138 \log_{10} \frac{2\phi}{d_0}, \tag{6}$$

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[1] D. E. Mode, "Low-pass filters using coaxial transmission lines as elements," Proc. I.R.E., vol. 36, pp. 1376–1383; November, 1948.

[2] Ramo and Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, Inc., New York, N. Y., chaps. 8 and 9; 1944.

[3] Radio Research Laboratory Staff, "Very High-Frequency Techniques," McGraw-Hill Book Company, Inc., New York, N. Y., chaps. 26, 27, and 32; 1947.

where d_1 is the diameter of a shunt rod. For completeness, three image impedance curves are shown in Fig. 4 which are typical of the band-pass structure of Fig. 2.

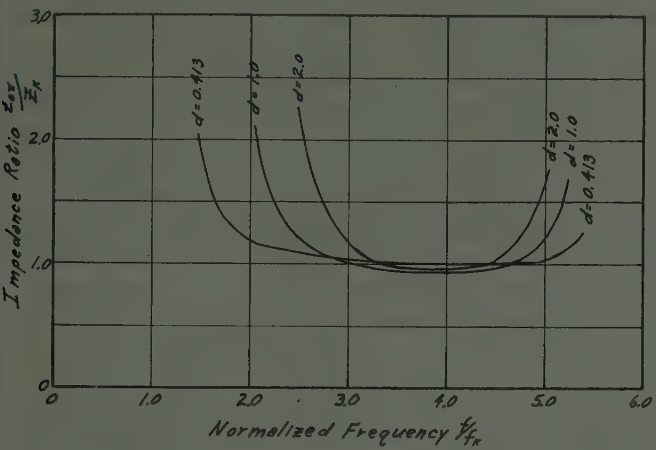

Fig. 4—Band-pass filter image impedance characteristics. $\phi=15°$, $\theta_1=26°$, at $f/f_k=1.0$.

III. Experimental Results

Twelve filter models were built and tested; these had in common the parameters $\phi=15°$, $\theta_1=26°$, and $f_k=1{,}155$ Mc. The filter image impedance was chosen to match a 52-ohm line at a value of $f/f_k=2.6$, giving the dimensions $D_k=2.74$ inches, $d_k=1.26$ inches, and $2\phi=0.854$ inch. Six filters had shunt rods of $\frac{1}{8}$ inch diameter and the other six shunt rods of $\frac{3}{8}$ inch diameter. Each of these filter groups involved the following:

A. A single shunt rod per section.
B. A single shunt rod per section, alternate rods transposed.
C. Two shunt rods per section, spaced 180°.
D. Two shunt rods per section, spaced 90°.
E. Two shunt rods per section, spaced 180°, transposed.
F. Four shunt rods per section, symmetrically spaced.

A photograph of a filter having four $\frac{3}{8}$-inch rods per section is shown in Fig. 5.

Fig. 5—Disassembled view of a band-pass filter having four shunt rods per midseries section.

An accurate way of computing either 2ϕ or d is not as yet evident. Experiments indicate that 2ϕ may be taken as the distance between the centers of shunt rods for a filter of type A, but should be measured between rod surfaces when the filter is type F. Also, the parameter d which is a measure of the shunt rod group impedance is not easily defined when a number of rods are used in a section. Although two rods spaced 180° apart may be assumed to have one-half the impedance of a single rod, the situation becomes more complicated due to mutual effects when either more rods are used or their spacing decreased.

The results of the tests on the twelve filter models are summarized in Table I. In describing the filter type, the first number indicates the number of shunt rods used per section, the letter T indicates transposition, and the final number specifies, in degrees, the separation between shunt rods in a given rod group. The second and third columns represent the experimental and theoretical lower TEM cutoff frequencies, respectively, and the fourth column lists the cutoff frequency of a spurious mode to be discussed in the next section.

TABLE I

Experimental and Calculated Lower Cutoff Frequencies of 12 Band-pass Filters

Filter Type	*TEM* Cutoff Frequencies in Mc		Calculated TE_{10} Cutoff Frequency in Mc
	Experimental	Calculated	
Filters With $\frac{1}{8}''$ Shunt Rods			
1	820	1,425	950
1-*T*	1,300	1,425	1,950
2–180°	1,800	1,875	1,950
2–90°	1,280	1,875	1,285
2-*T*-180°	1,940	1,875	4,075
4–90°	2,900	2,950	4,075
Filters With $\frac{3}{8}''$ Shunt Rods			
1	920	2,000	990
1-*T*	1,780	2,000	2,130
2-180°	2,000	2,500	2,130
2-90°	1,400	2,500	1,360
2-*T*-180°	2,600	2,500	4,920
4-90°	4,000	3,530	4,920

IV. Spurious Mode Transmission Past Shunt Rods in Coaxial Transmission Lines

Table I shows a poor agreement between the observed and calculated lower TEM cutoff frequencies in several cases. That this is due to spurious or higher order mode transmission is now to be shown.

Besides the principal or TEM wave there are also TM waves which may propagate in a coaxial transmission line; what is most important is that certain of these waves may travel past a shunt rod without attenuation due to the rod. The lowest order TM mode requires that there be room for about a half wavelength of field variation between the inner and outer conductors of the coaxial transmission line. This leads to

$$f_c = \frac{3 \times 10^{10}}{D_k - d_k}, \qquad (7)$$

where f_c is the lowest cutoff frequency (approximate) of the *TM* wave system and the dimensions are in centimeters. For the filters considered, (7) gives a *TM* cutoff frequency near 8,000 Mc and so a *TM* mode cannot be responsible for the anomalies turned up in Table I because they occur at much lower frequencies.

Certain *TE* modes are also able to propagate unattenuated past shunt rods. An approximate way of calculating the *TE* cutoff frequencies is to convert the coaxial transmission line into an equivalent rectangular waveguide by thinking of it as being rolled out flat so that the inner and outer conductors form the top and bottom surfaces of the hypothetical rectangular waveguide and the adjacent shunt rod surfaces become the sides. The width of the equivalent guide is the average distance between adjacent shunt rod surfaces. The *TE* cutoff frequencies may then be had from the well-known rectangular waveguide theory. A filter having but one shunt rod per section is shown in cross section in Fig. 6, in which the

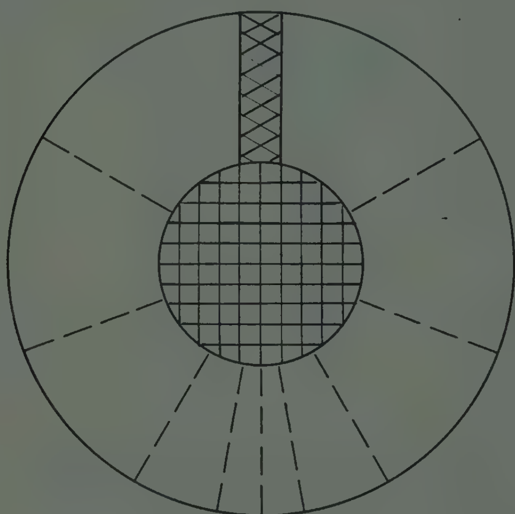

Fig. 6—Cross section of a filter having one shunt rod per section. The electric field is roughly indicated by the dashed lines.

electric field of the lowest order *TE* mode is shown by the dotted lines. This mode of transmission is termed a TE_{10} mode because there is but one cyclic variation in the electric field in the circumferential direction (between adjacent shunt rod surfaces) and none in the radial direction. The approximate cutoff frequency for this mode is,

$$f_c = \frac{3 \times 10^{10}}{\pi(D_k + d_k) - 2d_1} \tag{8}$$

with the dimensions in centimeters. Figs. 7 and 8 give rough cross-sectional views of filters having double shunt rods spaced 180° and 90°, respectively. Approximate equations for the TE_{10} cutoff frequencies are,

$$f_c = \frac{3 \times 10^{10}}{\frac{\pi}{2}(D_k + d_k) - 2d_1} \tag{9}$$

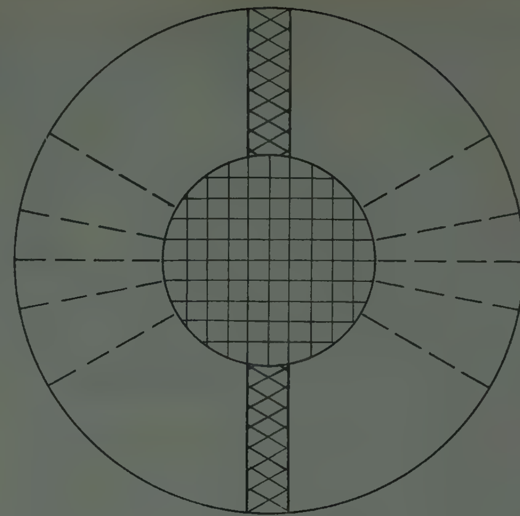

Fig. 7—Cross section of a filter having two shunt rods per section, spaced 180°. The electric field is roughly indicated by the dashed lines.

$$f_c = \frac{3 \times 10^{10}}{\frac{3\pi}{4}(D_k + d_k) - 2d_1} \tag{10}$$

for the filters of Figs. 7 and 8, respectively, and if four shunt rods are used, there results,

$$f_c = \frac{3 \times 10^{10}}{\frac{\pi}{4}(D_k + d_k) - 2d_1}. \tag{11}$$

The effect of transposing shunt rods is interesting. Consider, for instance, a filter using but one shunt rod per section and having these rods transposed. The transmission mode indicated in Fig. 6 could then propagate undisturbed only if the field could somehow twist through 180° in the small space between shunt rods. As

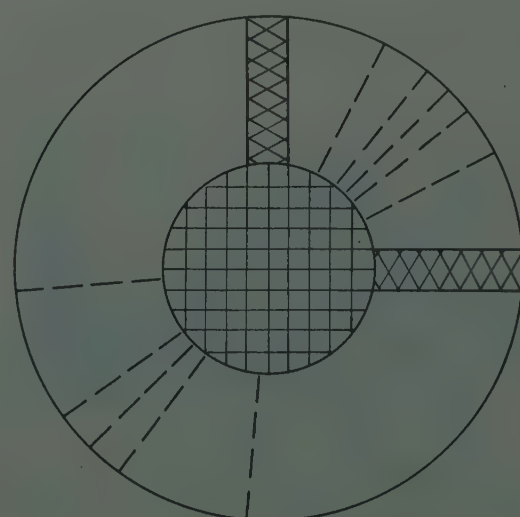

Fig. 8—Cross section of a filter having two shunt rods per section, spaced 90°. The electric field is roughly indicated by the dashed lines.

no reasonable mechanism exists which may account for such twisting, it might be expected that spurious mode transmission, if any, would have to be of the type shown in Fig. 7 for this field is able to propagate past the transposed rods without resort to rotation. Equation (9) should therefore be applicable to the case of single shunt transposed rods. Extensions of this idea should not be difficult for other combinations of transposed shunt rods.

The equations given above were employed to calculate the first order TE mode cutoff frequencies in the various filters reported, and the results are shown in the fourth column of Table I. It may be seen that in most cases where the agreement between experimental and theoretical TEM cutoff frequencies was poor, the inclusion of the possibility of spurious mode transmission explains the discrepancy. From the standpoint of design, the above indicates that if spurious mode transmission is to be eliminated in the useful portion of the attenuation band of a coaxial transmission line filter, a sufficient number of shunt rods must be used per section or else a transposition scheme employed so that spurious transmission occurs only at frequencies which are out of the working attenuation band of the filter.

Another spurious mode which is troublesome occurs when the circumference of either the inner or outer conductor of the coaxial line is equal to a multiple of a wavelength and a circumferential resonance may occur. This is not a new phenomenon, having been previously described by the Harvard research group.[3] Although circumferential resonances are not transmission modes, they are able to extract sufficient energy from the transmitted mode to introduce a "hole" in the pass band of a filter. For instance, in the filter models described, attenuation peaks of as much as 40 db were found in the neighborhood of 2,950 Mc. These were due to circumferential resonances on the inner conductor of the coaxial line (the circumference is 10.07 cm, corresponding to a full wave resonance frequency of 2,980 Mc). Slots cut axially into this conductor and filled with pencil lead (an absorbing material) substantially eliminated the attenuation peaks. In other filter designs not reported here resonances have been observed on the inside surface of the outer coaxial line conductor.

V. A More Exact Method for Calculating the TE Mode Cutoff Frequencies in Coaxial Transmission Lines Having Shunt Obstacles

The equations for the electromagnetic field of a TE wave in cylindrical co-ordinates are well known. For the problem at hand it is only necessary that certain boundary conditions be satisfied. At the surface of a shunt rod the radial component of the electric field must be zero, and at the surface of either the inner or outer conductor of the series transmission line, the electric field in the ϕ direction must be zero. The equations for these field components are,[2]

$$E_r = g(f, r)\left[AJ_n(k_c r) + B\genfrac{}{}{0pt}{}{N_n}{J_{-n}}(k_c r)\right] \cdot [-nC \sin n\phi + nD \cos n\phi] \quad (12)$$

$$E_\phi = h(f, r)\left[AJ_n'(k_c r) + B\genfrac{}{}{0pt}{}{N'_n}{J'_{-n}}(k_c r)\right] \cdot [C \cos n\phi + D \sin n\phi] \quad (13)$$

in which J_n and N_n are bessel functions of the first and second kinds. The choice of the J_{-n} or N_n function depends upon whether n is or is not fractional. The boundary conditions for E_r may then be satisfied if D is set equal to zero, the origin of ϕ chosen at the surface of a shunt rod, and the sine function made to vanish when $\phi = \psi$, where ψ is the angle between adjacent shunt rod surfaces. Because ψ is a function of the radius at which it is measured, the analysis is approximated by computing it at the mean radius. If fan-shaped shunt rods are used (true conical sections) this approximation becomes exact. The above gives,

$$n = \frac{m\pi}{\psi}. \quad (14)$$

If n is not an integer, which is usual, the lowest order TE mode has for its circumferential component,

$$E_\phi = h(f, r)[AJ'_{\pi/\psi}(k_c r) + BJ'_{-\pi/\psi}(k_c r)]C \cos \frac{\pi\phi}{\psi}, \quad (15)$$

which must vanish at the surfaces of the series transmission line conductors. Hence,

$$\begin{aligned} AJ'_{\pi/\psi}(k_c a) + BJ'_{-\pi/\psi}(k_c a) &= 0 \\ AJ'_{\pi/\psi}(k_c b) + BJ'_{-\pi/\psi}(k_c b) &= 0, \end{aligned} \quad (16)$$

where a and b are the radii of the inner and outer conductors of the series transmission line, respectively. The cutoff frequency in terms of k_c is,[2]

$$f_c = \frac{3 \times 10^{10} \times k_c}{2\pi}. \quad (17)$$

if the dimensions are in centimeters. The value of k_c is obtained by simultaneous solution of equations (16), the smallest value corresponding to the lowest order mode. As equations (16) are transcendental, a graphical solution must be resorted to, a better form for this purpose being,

$$\frac{J'_{\pi/\psi}(k_c a)}{J'_{-\pi/\psi}(k_c a)} = \frac{J'_{\pi/\psi}(k_c b)}{J'_{-\pi/\psi}(k_c b)}. \quad (18)$$

The process of finding a solution of (18) requires the computation of fractional order bessel function derivatives. The following relations are useful in this respect:

$$J_n'(x) = \tfrac{1}{2}[J_{n-1}(x) - J_{n+1}(x)] \quad (19)$$

and,

$$J_n(x) = \frac{1}{n!}\left(\frac{x}{2}\right)^n - \frac{1}{1!(n+1)!}\left(\frac{x}{2}\right)^{n+2} + \frac{1}{2!(n+2)!}\left(\frac{x}{2}\right)^{n+4} - \text{etc.}, \quad (20)$$

in which n is unrestricted and the factorials are had by reference to their definition in terms of the gamma function.

As an illustration, suppose that the cutoff frequency of the lowest TE mode in the filter having $\frac{3}{8}$-inch rods in 90° section be calculated. The mean value of ψ is 1.38π. Either side of equation (18) then becomes,

$$\frac{J'_{0.725}(x)}{J'_{-0.725}(x)} = \frac{J_{-0.275}(x) - J_{1.725}(x)}{J_{-1.725}(x) - J_{0.275}(x)}, \tag{21}$$

and a plot of this is shown in Fig. 9. The work is finished when the first two equal values of the function are found which are separated in x by the factor b/a. The dotted line in Fig. 9 defines two such values separated by the factor $b/a = 2.17$. As x is equal to about 0.47 at the first intersection, substitution into equation (17) gives 1,404 Mc for the cutoff frequency. This may be compared with the approximate theoretical value of 1,360 Mc and the experimental value of 1,400 Mc.

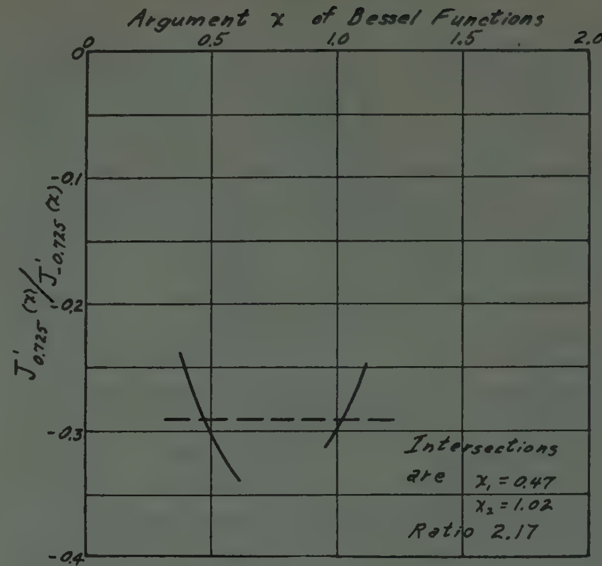

Fig. 9—Plot of equation (21) versus x. The dashed line defines values of x separated by the radius ratio b/a.

The Analysis of Broad-Band Microwave Ladder Networks*

M. C. PEASE†, MEMBER, IRE

Summary—By expressing the transmission matrix in terms of the Pauli spin matrices it has been found possible to write formulas for symmetrical ladder networks that are practical for analysis, even in the presence of gross line effect or parameter variation. These formulas are particularly useful for broad-band microwave filters.

General formulas are given for the combination of 2, 3, 4, and 5 elements. Detailed formulas are also given for elements that may be described as parallel resonant structures in shunt at the center of a length of line or waveguide.

An example is given illustrating the "line effect" in a low-Q structure that otherwise would show a "Butterworth" or "semi-infinite slope" characteristic.

Introduction

It is the purpose of this paper to present formulas for the analysis of ladder networks that appear to have considerable advantages when applied to broad-band microwave structures.

In general the analysis of wide-band microwave filters is greatly complicated by the existence of line effects. While normal methods[1,2] are theoretically applicable, the matrices and equations become hopelessly involved as the number of elements increases.

We shall give general formulas for symmetrical 2-, 3-, 4-, and 5-element structures which are practical, even in the presence of gross line effects or significant variation of the other parameters. It would not be difficult to extend these further, if desired. We shall give explicit formulas for quarter-wave coupled filters including the line parameter. Finally, we shall give an example of a 5-element filter with approximately a 13 per cent bandwidth. The derivation of the formulas is outlined in the Appendix.

Theory

We are concerned with filters constructed symmetrically by the series connection of known elements. Each element may be described by its transmission matrix,[1,2] relating its input voltage and current to its output. We shall write the matrix

$$((a)) = \begin{Vmatrix} A & jB \\ jC & D \end{Vmatrix}. \tag{1}$$

The matrix of successive elements, connected in series, is obtained by multiplying[2] the component matrices.

* Decimal classification: R143. Original manuscript received by the Institute, April 14, 1949; revised manuscript received, August 25, 1949.
† Sylvania Electric Products Inc., Boston, Mass.
[1] E. A. Guillemin, "Communication Networks," vol. II, John Wiley & Sons, Inc., New York, N. Y., 1947.
[2] P. I. Richards, "Applications of matrix algebra to filter theory," Proc. I.R.E., vol. 34, pp. 145 P–151 P; March, 1946.

As discussed in the Appendix, we define four coefficients for (1) by

$$a_0 = \tfrac{1}{2}(A + D) \qquad a_2 = \tfrac{1}{2}(-B + C)$$
$$a_1 = \tfrac{1}{2}(B + C) \qquad a_3 = \tfrac{1}{2}(A - D). \tag{2}$$

If the structure represented by ((a)) is lossless, these coefficients are all real. If the network is symmetric, a_3 vanishes at all frequencies.

If the network is normalized and matched, the voltage transmission function T = (voltage out)/(voltage in) is given by

$$\left|\frac{1}{T}\right|^2 = 1 + a_2^2 \tag{3}$$

for the symmetric case.

T is related to the VSWR (voltage standing-wave ratio), σ, by

$$\left|\frac{1}{T}\right|^2 = 1 + \frac{(\sigma - 1)^2}{4\sigma}\cdot \tag{4}$$

It is therefore necessary to calculate only a_2 for a symmetric network to obtain its significant behavior.

The essential formulas for the combination of these coefficients are given below. Double parentheses indicate matrices, while single parenthesis refers to the coefficient indicated by the subscript. a, b, c indicate successive elements in the network which are assumed to be individually symmetric (i.e., identically tuned).

Two Elements:

$$((a - a)) = ((a))\cdot((a))$$
$$(a - a)_0 = 2a_0^2 - 1$$
$$(a - a)_1 = 2a_0a_1$$
$$(a - a)_2 = 2a_0a_2 \tag{5}$$
$$(a - a)_3 = 0.$$

Three Elements:

$$((b - a - b)) = ((b))\cdot((a))\cdot((b))$$
$$(b - a - b)_0 = -a_0 + 2b_0g_{ab}$$
$$(b - a - b)_1 = a_1 + 2b_1g_{ab}$$
$$(b - a - b)_2 = a_2 + 2b_2g_{ab} \tag{6}$$
$$(b - a - b)_3 = 0$$

where

$$g_{ab} = a_0b_0 - a_1b_1 + a_2b_2. \tag{7}$$

Four Elements:

$$((b - a - a - b)) = ((b))\cdot((a - a))\cdot((b))$$

and the significant coefficient is

$$(b - a - a - b)_2 = 2a_0a_2 + 2b_2g_{aa-b} \tag{8}$$

where

$$g_{aa-b} = -b_0 + 2a_0g_{ab}. \tag{9}$$

Five Elements:

$$((c - b - a - b - c)) = ((c))\cdot((b - a - b))\cdot((c))$$
$$(c - b - a - b - c)_2 = a_2 + 2b_2g_{ab} + 2c_2g_{ab-c} \tag{10}$$

where

$$g_{ab-c} = g_{ac} + 2g_{ab}g_{bc} - 2a_0c_0. \tag{11}$$

The Microwave Filter

We shall apply these formulas to symmetrical filters of the type described by Fano and Lawson.[3] We shall assume that we may represent the structure as simple shunt elements identically tuned and spaced uniformly along a transmission line or waveguide. The method could easily be adapted to other structures, such as series elements, or mixed, or to varying line lengths.

If the doubly loaded Q[4] of the ((a)) element is denoted by "a," etc., and we include lengths of line $\theta/2$ electrical degrees on each side in our element, then, as may be found by normal methods,[1,2]

$$(a)_0 = \cos\theta - ax\sin\theta = \phi$$
$$(a)_1 = \sin\theta + ax\cos\theta \tag{12}$$
$$(a)_2 = ax$$

where

$$x = (f/f_0 - f_0/f) \tag{13}$$

and f_0 is the resonant frequency. More generally, the shunt susceptance of the element is ax, where x is any appropriate frequency parameter and a is independent of frequency, if possible, and contains the variation from one element to another.

Other elements will be denoted by b, c, etc., and will have the same coefficients, with b, c, etc., replacing a. The formulas for the transmission functions may be calculted from (5) through (11):

Two Elements: $((a-a))$

$$\left|\frac{1}{T_2}\right|^2 = 1 + 4(ax)^2\phi^2. \tag{14}$$[5]

Three Elements: $((b-a-b))$

$$\left|\frac{1}{T_3}\right|^2 = 1 + x^2(a - 2b + 4b\phi\psi)^2 \tag{15}$$

[3] R. M. Fano and A. W. Lawson, Jr., "Microwave filters using quarter-wave couplings," Proc. I.R.E., vol. 35, pp. 1318–1323; November, 1947.

[4] M. C. Pease, "Q measurements—two- and four-terminal networks," Proc. I.R.E., vol. 37, pp. 573–577; May, 1949.

[5] This is the simplest case of a formula given by M. C. Pease in a letter to the Editor, "A generalized formula for recurrent filters," to be published in Proc. I.R.E.

For n identical elements of this type, we have

$$\left|\frac{1}{T_n}\right|^2 = 1 + (ax)^2U_n^2(\phi)$$

where $U_n(\phi)$ is the Tschebyscheff polynomial of the second kind and order n in the variable ϕ.

where

$$\psi = \cos\theta - bx\sin\theta. \tag{16}$$

Four Elements: $((b-a-a-b))$

$$\left|\frac{1}{T_4}\right|^2 = 1 + \{2ax\phi - 2bx(\psi + 2\phi - 4\phi^2\psi)\}^2. \tag{17}$$

Five Elements: $((c-b-a-b-c))$

$$\left|\frac{1}{T_5}\right|^2 = 1 + \{(a-2c)x + 2bx(2\phi\psi - 1) + 4cx(2\phi\psi - 1)(2\psi\omega - 1)\}^2 \tag{18}$$

where

$$\omega = \cos\theta - cx\sin\theta. \tag{19}$$

Example:

In Fig. 1 we have computed the theoretical "exact" solution for a 5-element design that, in the absence of

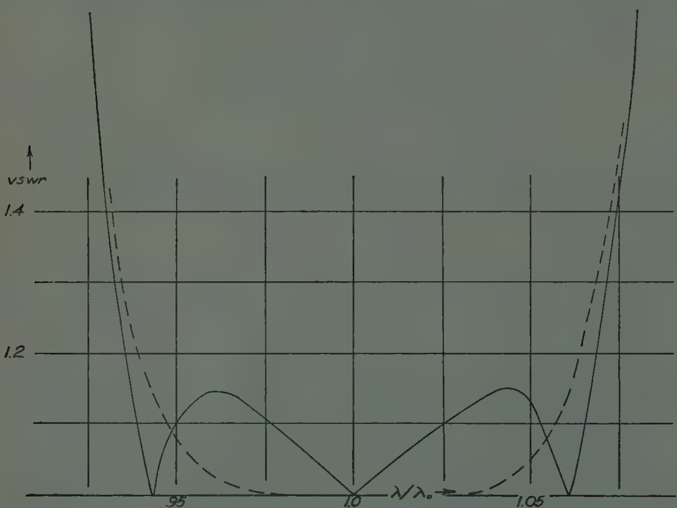

Fig. 1—"Exact" VSWR of 5-element filter which, without line effect, would give dotted line ("Butterworth" or Semi-Infinite Slope").

line effect, would give a "Butterworth," or "semi-infinite slope" solution. The doubly loaded Q's of the elements have been taken as

$$a = 4.94 \quad \text{(center element)}$$
$$b = 4.00 \quad \text{(second and fourth)}$$
$$c = 1.528 \quad \text{(outer elements).}$$

We have assumed that, if

$$y = f_0/f, \tag{20}$$

so that, from (13)

$$x = (1/y - y) \tag{21}$$

then

$$\theta = 90°/y. \tag{22}$$

From the x and θ calculated for a given value of y, ϕ, ψ, and ω are calculated from (12), (16), and (19).

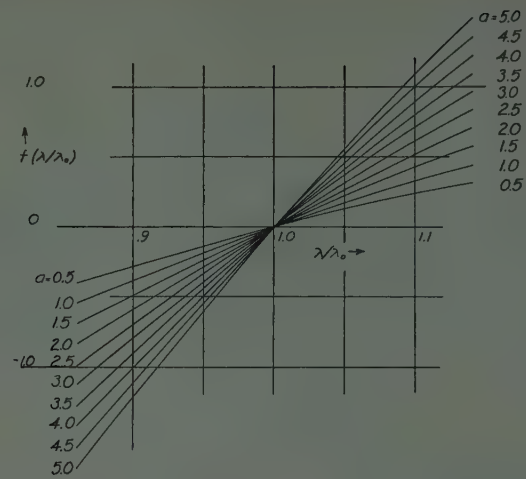

Fig. 2—$f(\lambda/\lambda_0)$ = "Reduced Susceptance" versus (λ/λ_0) for quarter-wave spacing for various values of doubly loaded Q, a.

$|1/T_5|^2$ is calculated from (18), and VSWR from (4).

It is interesting that the line effect actually widens the band, slightly, though with the introduction of "humps."

Curves

To simplify calculation, in Fig. 2 we have plotted, for various values of a, what we call the "reduced susceptance"

$$f(y) = \cos\theta - ax\sin\theta = \psi, \phi, \omega, \text{etc.,}$$

where θ has the frequency dependency of (20) and (22)—i.e., quarter-wave spacing, no waveguide effect (which may usually be neglected anyway).

In Figs. 3 and 4, we have replotted this data for various values of y as functions of a. Fig. 3 is for $y>1$, with $f(y)>0$. Fig. 4 is for $y<1$ with $f(y)<0$. This is usually the most convenient plot.

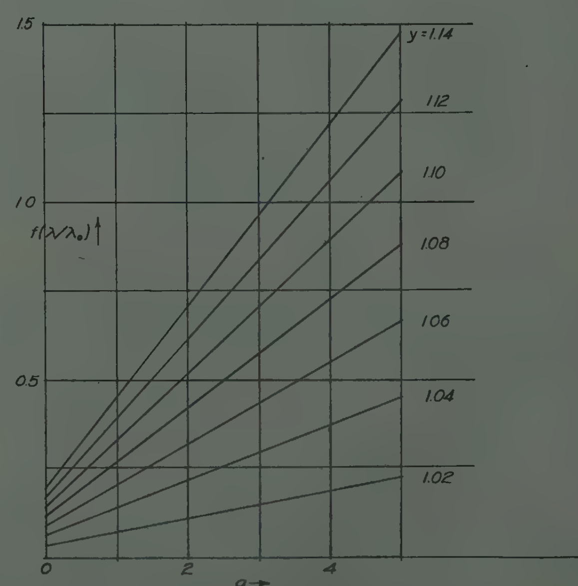

Fig. 3—$f(\lambda/\lambda_0)$ versus doubly loaded Q, a, for various values of $\lambda/\lambda_0>1$ ($\lambda_0/4$ spacing).

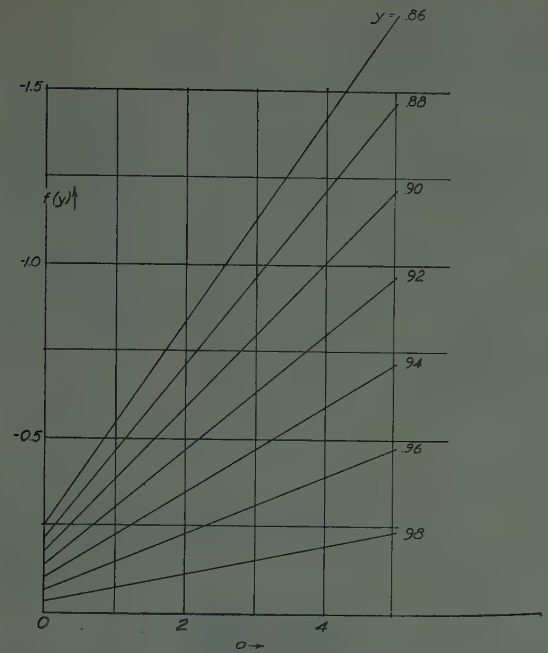

Fig. 4—$f(\lambda/\lambda_0)$ versus doubly loaded Q, a, for various values of $\lambda/\lambda_0<1(\lambda_0/4$ spacing).

Conclusion

We have given formulas for the exact calculation of the transmission function of a symmetrical ladder structure. These are in a form suitable to low-Q microwave filters. We have also given the detailed formulas for elements which are identically tuned units equivalent to parallel resonant networks placed in shunt to a line or waveguide at equal intervals.

We have, as an example, calculated the exact VSWR for the sequence of elements that, in the absence of a "line effect," gives rise to the "Butterworth" or "semi-infinite slope" characteristic. The line effect is pronounced, introducing very appreciable "humps" in the passband.

Finally, we give curves of the significant function, $f(y)=\phi$, ψ, ω, etc., for the commonest case of center-band quarter-wave spacing. This permits the ready calculation of the VSWR of a low-Q filter, including line effect.

The theory is presented in the Appendix. The device used is to expand the transmission matrix as functions of a group which are essentially the Pauli spin matrices. This representation has certain very significant advantages.

Appendix

The coefficients (2) are obtained by writing

$$((a)) = a_0 I + a_1\alpha + a_2\beta + a_3\gamma \qquad (23)$$

where I, α, β, and γ are the 2-square matrices:

$$I = \begin{Vmatrix} 1 & 0 \\ 0 & 1 \end{Vmatrix} \qquad \alpha = \begin{Vmatrix} 0 & j \\ j & 0 \end{Vmatrix}$$
$$\beta = \begin{Vmatrix} 0 & -j \\ j & 0 \end{Vmatrix} \qquad \gamma = \begin{Vmatrix} 1 & 0 \\ 0 & -1 \end{Vmatrix}. \qquad (24)$$

I, α, β, and γ may be recognized as the Pauli spin matrices[6] (except for a factor of j in α). They form a group having the multiplication table

$$\begin{array}{cc|ccc} & \multicolumn{4}{c}{\textit{2nd term}} \\ & I & \alpha & \beta & \gamma \\ \hline & \alpha & -I & -\gamma & \beta \\ \textit{1st term} & \beta & \gamma & I & \alpha \\ & \gamma & -\beta & -\alpha & I \end{array} \qquad (25)$$

One of their important advantages is that they are anti-commuting—i.e., for example, $\alpha\beta=-\beta\alpha$.

The elements of the original matrix (1) are related

$$AD + BC = 1. \qquad (26)$$

Hence, from (2), the coefficients are related

$$a_0^2 + a_1^2 - a_2^2 - a_3^2 = 1. \qquad (27)$$

It is of some interest, perhaps, that (27) indicates we may consider $((a))$ as a unit space vector in a four-dimensional metric of which two dimensions are "space-like" and two, "time-like." In the case of symmetric filters, where $a_3=0$, the metric reduces to the world metric of Minkowski for relativisitic space.[7]

(3), the equation for $|1/T|^2$, is obtained from

$$|1/T|^2 = \tfrac{1}{4}|A + D + j(B + C)|^2 \qquad (28)^{1,2}$$

by assuming the network is lossless, so that A, B, C, and D are real and

$$|1/T|^2 = \tfrac{1}{4}(A + D)^2 + \tfrac{1}{4}(B + C)^2 \qquad (29)$$

$$= a_0^2 + a_1^2 \qquad (30)$$

from (2). Hence, by (27)

$$|1/T|^2 = 1 + a_2^2 + a_3^2. \qquad (31)$$

In the symmetrical case where $a_3=0$, (31) reduces to (3).

The multiplications leading to (5) and (6) and their further expansion to (8) and (10) may be easily done with the aid of (25), using (27) where appropriate.

The advantages of this somewhat devious method are twofold. The anticommuting property leads to the cancellation of many of the cross products in, for example, the calculation of $((b-a-b))$. Furthermore, the extremely simple form of (3) means that only a single coefficient of a symmetric combination need be found.

[6] V. Rojansky, "Introductory Quantum Mechanics," Prentice-Hall, Inc., New York, N. Y., p. 482, et seq; 1946.

[7] P. G. Bergmann, "Introduction to the Theory of Relativity," Prentice-Hall, Inc., New York, N. Y., 1947.

Contributors to the Proceedings of the I.R.E.

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N. R. Castellini

Since 1940 Mr. Castellini has been employed at the Signal Corps Engineering Laboratories, where he has been associated in the design of army ground radio communication equipment.

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John P. Day (M'47) was born on January 17, 1919, at David City, Neb. In 1940 he received the B.S. degree in physics from the California Institute of Technology; he also studied physics in the Graduate School of the same institution during 1940 and 1941. From 1941 to 1943 he was engaged in aircraft radar development at the Naval Research Laboratory, in Washington, D.C. Since 1943 he has been employed in the Research Division of the United States Navy Electronics Laboratory in San Diego, Calif.

John P. Day

❖

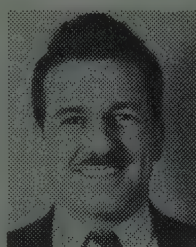
J. Z. Millar (A'30–SM'45) was born in Mattoon, Ill., on July 3, 1901. Following his graduation from the University of Illinois in 1923, he joined the Western Union Telegraph Company as an engineering apprentice, transferring in 1926 to the Water Mill laboratory to specialize in electronics. After doing research on short-wave equipment and audio-frequency apparatus for fifteen years, Mr. Millar was called to active duty with the Signal Corps, in which he attained the rank of colonel. He served as member and director of the Signal Corps Board in Ft. Monmouth, N. J., from March, 1941, to April, 1944, and was then assigned as Signal Officer of the Normandy Base section and as Signal Officer, Loire Section, European Theater.

J. Z. Millar

On February 15, 1945, he returned to Western Union and was appointed radio research engineer. In this position he has organized the radio research division of the development and research department, which division handles all radio research projects. On August 1, 1949, Mr. Millar was appointed director of research.

❖

M. L. Miller was born in Indiana on January 23, 1916. He received the B.S.E.E. degree from Purdue University in 1939 and worked for the Indiana Michigan Electric Company until September, 1941, when he entered the U. S. Army Signal Corps. He served in numerous commands in the ETO and the United States as radar officer, communications officer, or instructor during the following four and one-half years. Upon leaving the Signal Corps, he joined the Capehart-Farnsworth Corporation as an electronics engineer. Mr. Miller is a member of the American Institute of Electrical Engineers.



M. L. Miller

❖

Charles W. Mueller (S'35–A'36–VA'39–SM'45) was born in New Athens, Ill., in 1912. He received the B.S. degree in electrical engineering from the University of Notre Dame in 1934, and the S.M. degree in electrical engineering from the Massachusetts Institute of Technology in 1936. From 1936 to 1938 Dr. Mueller was associated with the Raytheon Production Corporation, first in the engineering supervision of factory production of receiving tubes, and then in the development of gas-tube voltage regulators and cold-cathode thyratrons. In the fall of 1938 he returned to the Massachusetts Institute of Technology where he continued his studies and received the degree of Sc.D. in physics in 1942. While at the Massachusetts Institute of Technology, he worked on a government contract on the development of gas-filled special-purpose tubes for counting operations. From 1942 to the present he has been a member of the RCA Research Laboratories, where he has been engaged in research on high-frequency receiving tubes and secondary electron emission phenomena.

C. W. Mueller

Dr. Mueller is a member of the American Physical Society and Sigma Xi. He was the 1946–1947 Chairman of the Princeton Section of the IRE.

❖

George Sinclair (A'37–SM'46) was born in Hamilton, Ontario, Canada, on November 5, 1912. He received the B.Sc. degree in electrical engineering in 1933 and the M.Sc. degree in 1935 from the University of Alberta, followed by the Ph.D. degree in 1946 from the Ohio State University. Dr. Sinclair was an instructor in electrical engineering at the University of Alberta for one year, and engineer for the Northern Broadcasting Corporation for two years.

George Sinclair

From 1941 to 1947 Dr. Sinclair was a research associate in the department of electrical engineering of the Ohio State University, supervising the research program of the Antenna Laboratory. He is now assistant professor of electrical engineering at the University of Toronto.

❖

W. B. Sullinger was born in Fredericksburg, Va., on September 12, 1908. He joined the staff of the research division of the Western Union Telegraph Company following his graduation from the Virginia Polytechnic Institute in 1930. During the next several years he participated in studies involving direct-current telegraph transmission, interference mitigation, and carrier transmission. He later assisted in the design, development, and field testing of the carrier systems now widely used by Western Union. Since 1944, Mr. Sullinger has been concerned principally with the development of microwave relay systems for telegraph and television. He was appointed assistant radio research engineer in 1945, and radio research engineer on August 1, 1949.

W. B. Sullinger

C. C. Wang (A'43–SM'49) was born in China, on October 20, 1914. He received the B.S. degree in electrical engineering from Chiao-Tung University, in Shanghai, China, in 1936. During the period 1937–1940 he studied radio communication at Harvard University under the Tsing-Hua University fellowship. He received the S.M. and Sc.D. degrees in radio communication from Harvard in 1938 and 1940, respectively. After a year of study in radio tube manufacturing at the RCA Manufacturing Company, in Harrison, N. J., he was employed as a research engineer in the electronic engineering department of the Westinghouse Electric Corporation, in Bloomfield, N. J., remaining until 1945.

C. C. Wang

Dr. Wang's principal interests during the years 1942–1945 were concerned with research and development of microwave tubes. Since 1946 he has been affiliated with the electron tube department of the Sperry Gyroscope Company, Great Neck, L. I., N. Y., engaged in similar research on similar problems. He is a member of Phi Tau Phi, and of the American Association for the Advancement of Science.

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For a photograph and biography of Douglas E. Mode, see page 90 of the January, 1950, issue, of the Proceedings of the I.R.E.

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For a photograph and biography of M. C. Pease, see page 578 of the May, 1949, issue of the Proceedings of the I.R.E.

L. G. Trolese (A'42–SM'47) was born on July 2, 1908, in Sonora, Calif. He received the B.S. degree in electrical engineering at the University of California in 1930. His engineering experience includes one year with RCA Victor Company as a field engineer, and five years with General Engineering and Heating Company of San Francisco as electrical engineer. Since 1942 he has been with the U. S. Naval Electronics Laboratory at San Diego, Calif., where he has been engaged in electromagnetic wave-propagation studies. Mr. Trolese is currently serving on the IRE Paper Procurement Committee, and is Chairman of the San Diego Section of the IRE.

L. G. Trolese

Correspondence

"Weighted Average of Voltage" Method*

In a linear passive network which may be diagrammed below, where the two indicated matrices are the normal transmission matrices of their respective subnetworks, and it is desired to find the transient voltage E_3 under excitation by E_1, the following method has been found extremely useful and time-saving.

Designate the Laplace transform of a quantity $(\ldots)$ by $L(\ldots)$, and its inverse transform by $L^{-1}(\ldots)$, and use p as the variable in the transform. Further, write the transform of a voltage or current—symbolized by large Roman letters—by the respective small letters.

With reference to the diagram:

$$\left\| \begin{matrix} E_1 \\ I_1 \end{matrix} \right\| = \left\| \begin{matrix} F & H \\ K & N \end{matrix} \right\| \cdot \left\| \begin{matrix} E_3 \\ I_3 \end{matrix} \right\| \quad \begin{matrix} F=\alpha A+\beta G \\ H=\alpha B+\beta M \\ K=\gamma A+\mu G \\ N=\gamma B+\mu M \end{matrix}$$

and if

$$L[Z(j\omega)] = z(p)$$

* Received by the Institute, October 13, 1949.

then

$$e_3 = z(p) i_3$$

where $z(p)$ and (see below) $n(p)$, $r(p)$, are rational algebraic functions of p.

By transforming the above matrix expression we get

$$\begin{aligned} e_1 &= L(FE_3) + L(HI_3) \\ &= L\{FL^{-1}[z(p)i_3]\} + L(HI_3) =, \\ &\qquad \text{say, } n(p)i_3. \end{aligned}$$

Similarly

$$\begin{aligned} e_2 &= L(AE_3) + L(BI_3) \\ &= L\{AL^{-1}[z(p)i_3]\} + L(BI_3) =, \\ &\qquad \text{say, } r(p)i_3; \end{aligned}$$

then

$$E_2 = L^{-1}\left[\frac{r(p)}{n(p)} e_1\right].$$

Initial conditions may readily be taken account of in performing the transformations.

This method of finding the "weighted average of voltage" in many cases permits a great reduction in time and background computations. The circuit in the diagram has been so drawn because this method was originally developed for analyzing transients in cascade or transmission line structures.

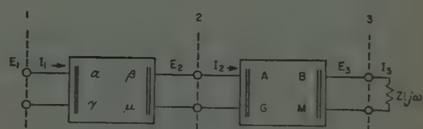

Fig. 1.

It should be noted that which element constitutes the termination of the network is often a matter of choice, so that even if geometrically the original network and transient are not positioned to resemble the diagram, this method is applicable if some topological distortion of the network would bring it into a form equivalent to that pictured.

A. A. Grometstein
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Boston, Mass.

Institute News and Radio Notes

Technical Committee Notes

The Standards Committee at a meeting on November 10, under the Chairmanship of Professor J. G. Brainerd, **approved all but three of the Proposed Electron Tubes Method of Testing.** Professor Brainerd announced that the Electronic Computers Committee has compiled a bibliography of published material in its field. Axel Jensen, Vice-Chairman of the Standards Committee, reported for the Definitions Co-ordinating Subcommittee. Work in the Transducer Task Group has started under the Chairmanship of Jack Morton. C. J. Hirsch has started work as Chairman of the Task Group on Pulse Definitions. The IRE AIEE Joint Committee on Noise Definitions does not yet have a Chairman. M. W. Baldwin reported activities of the Television Co-ordinating Subcommittee. . . . The first **meeting of the Subcommittee on Theory and Application of Tropospheric Propagation** was held on November 3, at the National Bureau of Standards, Washington, D. C. The Chairman is Dr. H. G. Booker of Cornell University. The fundamental purpose of this Committee is to simplify calculations and to assist in locating quickly the most useful sources of information. The final product will probably be a bibliography which will be published in the PROCEEDINGS OF THE I.R.E. . . . **The Board of Directors of the IRE** on November 16, **approved the change in the names of the following committees:** The Railroad and Vehicular Communications Committee will be known as the Mobile Communications Committee. The name of the Instruments and Measurements Committee has been changed to the Committee on Measurements and Instrumentation . . . . **The Modulation Systems Committee** held a meeting on November 17, under the Chairmanship of H. S. Black, to consider a list of proposed modulation definitions which was submitted by the Subcommittee on PCM Theory under date of November 8. . . **The Electron Tubes Committee** virtually completed work on Electron Tubes Definitions at a meeting November 16. . . **The Committee on Measurements and Instrumentation** held a meeting on November 18, under the Chairmanship of Professor Ernst Weber. The Chairman listed the Subcommittees as follows: Basic Standard and Calibration, Chairman, to be announced; Dielectric Measurements Subcommittee, J. Dalke, Chairman; Magnetic Measurements, Chairman to be announced; Audio-Frequency Measurements Subcommittee, A. P. G. Peterson, Chairman; Video-Frequency Measurements, Chairman to be announced; Radio-Frequency Measurements, Chairman to be announced; Noise Measurements Subcommittee, Chairman to be announced; Interference Measurements Subcommittee, K. A. Chittick, Chairman; Oscillography Subcommittee, P. S. Christaldi, Chairman; Basic Techniques of Instrumentation Subcommittee, Norman Taylor, Chairman; Telemetering Subcommittee, W. G. Mayo-Wells, Chairman; Electronic Components Subcommittee, J. G. Reid, Chairman. These Subcommittees will be working groups engaged in the preparation of standards. The subject of a Symposium on Basic Circuit Elements at the IRE National Convention was discussed by the Committee. Dr. Weber stated that he had organized this Symposium to cover the performance of coils, resistors, capacitors, and transformers. Mr. Frick summarized his report on the FCC Incidental Radiation Devices Conference in Washington on November 1. He represented the IRE Industrial Electronics Committee, and the Committee on Measurements and Instrumentation. Mr. Chittick, who represented the Receivers Committee, also reported briefly on his phase of the FCC Conference. Chairman Weber reported that the Symposium on Packaged Electronics to be sponsored jointly by IRE and AIEE in May, 1950, was in the initial stages of preparation. This is another excellent example of joint activity. . . . **The Audio Techniques Committee meeting** was held on December 1, with Edward J. Content, Chairman. W. L. Black reported on the material to be submitted to the Annual Review Committee. . . . **The Sound Recording and Reproducing Committee** held a meeting on November 29, under the Chairmanship of S. J. Begun. This committee will present papers at the Symposium on Sound Recording to be held during the 1950 IRE Convention. It was suggested that sufficient time should be left after the presentation of each paper for round-table and open discussions. The following material is to be presented: A General Discussion of Noise, by F. L. Hopper; Definitions and Measurements of Noise, by A. W. Friend; Frequency Response Measurements, by R. E. Zenner; Distortion Measurements, by J. Z. Menard; Problem of Flutter and Wow, by Harry Schecter. In addition to the foregoing papers, Dr. Begun will present a paper summarizing the general philosophy of recording touching upon the various current methods in use. Chairman Begun introduced C. F. West, who will become a member of the committee to facilitate liaison with the Electronic Computers Committee. His committee's interest would overlap with this committee's mainly in regard to pulse recording. Close liaison with the Electronic Computers Committee will be needed when the latter committee defines specific parts of equipment. . . . **The Task Group on Pulse Definitions** held a meeting on December 6, under the Chairmanship of Charles J. Hirsch. This task group recently appointed by the Standards Committee will correlate all IRE definitions pertaining to pulses. A single set of pulse definitions will be submitted to the Standards Committee. . . . **The Joint Technical Advisory Committee** held a meeting on November 29, under the Chairmanship of Donald G. Fink. James Veatch, consultant to JTAC, reported on the FCC informal Conference on Incidental Radiation Devices. JTAC will have a representative in attendance at future FCC meetings on this subject. . . . At a recent **meeting of the Administrative Committee of the Audio Professional Group**, Leo L. Beranek was appointed Chairman to fill the unexpired term of O. L. Angevine. . . . **The Administrative Committee of the Antennas and Propagation Professional Group** held a meeting on November 2, at the Engineering Laboratories of Jansky and Bailey, Washington, D. C. A report of the **Membership Committee** was given by P. S. Carter, Chairman. Mr. Carter reported that there were 509 members of the Professional Group as of October 30, 1949. Dr. Van Atta has been appointed Chairman of the Meetings Committee. . . . **The Administrative Committee of the Professional Group on Quality Control** held a meeting on November 1, at Syracuse, N. Y., under the Chairmanship of R. F. Rollman. Plans were discussed for the IRE Annual Convention Symposium. . . . **The Administrative Committee of the Nuclear Science Professional Group** held a meeting on November 1, at the Hotel Commodore, New York City. The Constitution and By-Laws as originally submitted have been approved by the Executive Committee. Seventy applications for membership were submitted and approved. The Nuclear Science Group will sponsor a Symposium at the 1950 IRE National Convention. The tentative program includes: Elementary Particles, Urner Liddel; Particle Accelerators, author to be announced; RF Problems in Accelerator Designs; Methods of Radiation Detection, Dr. John R. Dunning. . . . **The Steering Committee of the Joint IRE AIEE Symposium on Packaged Electronics** held a meeting on November 21 under the Chairmanship of F. Given. This Symposium will consist of a two-day meeting which will be held in Washington, D. C. in the Spring, 1950. The exact dates, speakers, and papers will be announced later.

FMA-NAB Liaison Committee Maps Proposal for Merger

Members of the Board of Directors of the Frequency Modulation Association have voted to accept proposals for a contemplated merger with the National Association of Broadcasters. An FMA-NAB Liaison Committee mapped proposals for the merger which will combine the interests of the two organizations.

Recommendations of this joint group subsequently were approved by NAB's board of directors at a meeting in November, following which an invitation was extended to FMA to merge the two associations.

No official date was set for effecting the merger, but January 1 was the tentative target date, FMA-NAB officials said.

URSI/IRE Joint Meeting Held in Washington Last Fall

The Joint Fall Meeting of the International Scientific Radio Union and The Institute of Radio Engineers held in Washington, D. C., on October 31, November 1 and 2, 1949, was sponsored by the USA National Committee of the URSI and the recently organized Professional Group on Antennas and Wave Propagation of the IRE. The session was a departure from the practice followed in previous years of having a general meeting in the fall as in the spring.

Sponsoring URSI Commissions were as follows: 2. Tropospheric Radio Propagation, Chairman, Dr. Charles R. Burrows; 3. Ionospheric Radio Propagation, Chairman, Dr. Newbern Smith; 5. Extraterrestrial Radio Noise, Chairman, Dr. D. H. Menzel; 6. Radio Waves and Circuits, including General Theory and Antennas, Chairman, Dr. L. C. Van Atta.

Papers invited by the Chairmen of the Commissions, informal discussions, and a few contributed papers were heard at the technical sessions at the first two days of the meeting. On November 3 administrative meetings were held by the various Commissions sponsoring the session. In the afternoon a joint technical session of all the Commissions summarized the activities and closed the meeting. Three hundred and forty persons attended these sessions and many of them participated in the interesting and lively discussions which accompanied the papers.

Abstracts of the papers were prepared in booklet form as a program. Copies are available at a price of $1.00 each and may be obtained upon request from Dr. Newbern Smith, Secretary-Treasurer of the USA National Committee, Central Radio Propagation Laboratory, National Bureau of Standards, Washington 25, D. C.

Midwest Power Conference Is Scheduled for April in Chicago

"Economy in Power" is the theme of the twelfth annual Midwest Power Conference to be held April 5, 6, and 7, 1950, at the Sherman Hotel, Chicago, according to an announcement from Conference Director Roland A. Budenholzer, professor of mechanical engineering at the Illinois Institute of Technology. Dr. E. R. Whitehead, director of the electrical engineering department at Illinois Tech, is conference secretary.

The annual three-day meeting is sponsored by the Illinois Institute of Technology with the co-operation of 18 midwestern universities and professional societies. Attracting each year over 3,000 engineers from all parts of the United States and Canada, it is now the largest conference of its kind in the world.

Co-operating institutions are: Iowa State and Michigan State Colleges, Northwestern and Purdue Universities, and the Universities of Iowa, Illinois, Michigan, Minnesota and Wisconsin, Western Society of Engineers, National Association of Power Engineers, Engineers' Society of Milwaukee, the Illinois chapter of American Society of Heating and Ventilating Engineers, the Illinois section of American Society of Civil Engineers, and the Chicago sections of American Institute of Electrical Engineers, American Institute of Chemical Engineers, American Institute of Mining and Metallurgical Engineers, and American Society of Mechanical Engineers.

Rider Receives ESFETA Radio Education Award

John F. Rider, president of John F. Rider, Publisher, Inc., of 480 Canal Street, New York 13, N. Y., was the recipient of an award given by the Empire State Federation of Electronic Technicians Association at the banquet of the Radio Technicians' Guild of Rochester, N. Y., held at Locust Lawn, Ionia, N. Y.

He received the award for his unceasing efforts on behalf of the radio-television servicemen of the country, it was announced. Mr. Rider was instrumental in inaugurating the current ESFETA TV lecture series, having delivered the opening talk of the series. In addition, during the past year he has traveled extensively for ESFETA, lecturing at servicemen's meetings.

The author of numerous textbooks now being used by radio servicemen and technical educational institutes, Mr. Rider has participated in the educational development of the radio serviceman since 1921.

General Electric Co. Builds System for Microwave Relay

General Electric Company's Switchgear Divisions are now building a "push-button" remote control system, in which radio and wire signals will operate eight electric power distribution stations covering a 450-mile area, according to information from the GE Apparatus News Bureau. From one central point, the system will control initially 226 and ultimately 384 individual units of electric apparatus.

The supervisory control system will be installed in the Colton station of the Southern California Edison Company. From there a master control panel will switch on and off equipment in the outlying stations as needed by the power distribution network. In addition to controlling individual pieces of equipment, the supervisory control system will also report the readings on 28 instruments and 41 alarm indicators at the remote stations.

NAB Completes Its Internal Reorganization of Structure

The Board of Directors of the National Association of Broadcasters has completed internal reorganization of the Association, a task which was launched over a year ago.

Additionally, the Board approved a proposal recommending dissolution of the Broadcast Measurement Bureau, the audience measurement organization which is jointly directed by NAB, the American Association of Advertising Agencies, and the Association of National Advertisers. The Board has proposed forming a new corporation to take its place.

In other actions, the 25-man group, in its quarterly meeting extended an invitation to the FM Association to merge with the NAB, and rescinded a former action which had set forth a plan to separate the Broadcaster Advertising Bureau, now a department of the NAB, from the parent structure, agreeing that it should remain as a department of the Association.

The actions were taken principally upon the recommendations of the Board's Structure Committee which had been created last year to present a plan for realigning the functions of the Association and the structure of its internal organization.

First Electronic Exhibit Presented in New Mexico

More than 1,000 persons attended the first New Mexico Electronic Exhibit held November 4 and 5 at the Alvarado Hotel, Albuquerque, N. M. Meetings of the New Mexico Section of the IRE, and the AIEE's Albuquerque subsection of the El Paso Chapter, were held in connection with the display.

Thirteen exhibitors represented 81 manufacturers who displayed the latest electronic instruments, allied testing, and laboratory equipment.

Calendar of COMING EVENTS

AIEE Winter General Meeting, New York, N. Y., January 30–February 3, 1940

ASTM Meeting on Radio Tube Materials, Committee B4-VIII, Philadelphia, Pa., March 1–2

1950 IRE National Convention, New York, N. Y., March 6–9

IRE/URSI Symposium on Antennas and Propagation, San Diego, April 3–5

IRE/URSI Meeting, Commissions 1, 6, 7, 4, Washington, D. C., April 17–19

Fourth Annual Spring Technical Conference, Cincinnati Section, IRE, April 29, Cincinnati, Ohio

1950 IRE Technical Conference, Dayton, Ohio, May 3–5

Conference on Improved Quality Electronic Components, sponsored by IRE, AIEE, RMA, Washington, D. C., May 9, 10, and 11.

Armed Forces Communications Association 1950 Annual Meeting, May 12, Photographic Center, Astoria, L. I., N. Y., and New York City; May 13, Signal Corps Center, Fort Monmouth, N. J.

Industrial Engineering Notes[1]

Television News

Television receiver production in October rose another 35 per cent to reach a new peak in response to a pre-Christmas trade of boom proportions. RMA member companies reported the manufacture of 304,773 TV sets in a four-week October period—35 per cent more than in September, and total industry production was estimated at more than 360,000. Television receivers reported by RMA set manufacturers numbered 1,707,613 for ten months with total industry production estimated at more than 2,000,000 TV sets. Radio receiver production also rose in October to meet a revived market for both AM and FM sets. FM and FM-AM radios reported by RMA companies in October totalled 83,013, the highest mark since last February, with an additional 50,545 TV sets reported as equipped with FM reception facilities. Total set production—both TV and radio—of RMA members reached 975,053, the highest output for the year 1949. . . . The FCC has announced it will not grant any new authorizations for point-to-point television relays for the purpose of exhibitions or demonstrations of television programs, including large-screen theater television. The Commission noted that in recent weeks it has issued certain special temporary authorizations permitting the use of radio frequencies for the purpose of point-to-point relay of television programs, which were either exhibited on broadcast receivers or on large-screen theater television equipment. It was pointed out that these authorizations were approved because the parties were not acquainted with FCC rules and had made arrangements before seeking permission. The FCC notice pointed out that no frequencies have been allocated for such services, and that the Commission now has under consideration several petitions requesting the institution of rule-making proceedings, looking toward the establishment of theater television service and the allocation of frequencies for this service. . . . Sales of cathode-ray tubes for television receivers during the first nine months of 1949 were nearly double the value of such sales during the entire year of 1948, an RMA tabulation of manufacturer reports revealed.

Sales of TV receiver-type picture tubes totalled 2,129,210 units valued at $62,525,-446 in the first three quarters of 1949, compared with 1,309,176 units valued at $33,459,554 in the full year 1948. . . . Over the dissents of Chairman Wayne Coy and Commissioner George Sterling, the FCC yesterday scheduled a hearing to start January 16 before Commissioner Frieda Hennock on the petition of the Zenith Radio Corp. (*RMA Industry Report*, vol. 5, no. 31) for authority to conduct "Phonevision" tests on a limited commercial basis. Chairman Coy, with the concurrence of Commissioner Sterling, dissented with the FCC majority and said he favored granting the Zenith petition without a prior hearing for a three-month period of tests. Zenith, in its petition proposed to conduct experimental operation employing "Phonevision" on its experimental TV broadcast station W9XZV on Channel 2 on a limited commercial basis, involving 30 test subscribers who would be provided with TV sets equipped to receive "Phonevision" transmissions and who would be asked to pay specified fees for these programs. The FCC, in an order scheduling the hearing, said "It is not clear whether 'Phonevision,' as proposed by the petitioner, should be classified as a broadcast service, a common carrier, or other type of communications service, or what frequencies, if any, are appropriate for use in the proposed experimental operations or for use in a general commercial 'Phonevision' service in the event such service were to be authorized on a regular basis." . . . There was a total of 94 television stations on the air at the end of November and 17 others under construction, according to a recent check of FCC records. . . . As the result of petitions from RMA and CBS, the FCC has issued a new schedule for the resumption of the color television hearings and demonstrations in February. Under the revised schedule, the FCC postponed the initial demonstration by Color Television, Inc., of its proposed color system from February 6 to February 20. The date of the second comparative color TV demonstration by Columbia Broadcasting System, Inc., Color Television, Inc., and Radio Corporation of America was changed from February 8 to February 23–24. Because of these changes, the Commission postponed from February 13 to February 27, the resumption of additional direct testimony to be followed by cross examination of witnesses, including President R. C. Cosgrove. . . . Two steps in medical television were demonstrated in Washington at the third annual clinical session of the American Medical Association. One was the use of equipment in which the camera was actually located in the center of the operating room light. The other was another demonstration of color television at John Hopkins Hospital in Baltimore which was microwaved directly to receiving sets in the Armory in Washington. The color demonstration was staged with equipment developed by the Columbia Broadcasting System.

FCC Actions

The FCC has written to Arco Electronic Kit Division of New York City rejecting a proposal made recently by that concern for the testing of the proposed color television systems by radio amateurs or "experimenters." The Commission wrote that the advice of Arco was appreciated, but that it felt that the plan "would cause an indefinite and unnecessary delay in the final determination of the issues." . . . The FCC amended its notice of hearing concerning the use of facsimile broadcasting in connection with FM radio stations to include patent information on inventions relating to facsimile multiplex transmitters and receivers described in patent applications now pending. . . . **A temporary waiver and temporary rules concerning operator requirements for ship radar stations were extended** by the FCC from the present cutoff date of November 15, 1949, to May 15, 1950, or the effective date of permanent rules in this matter, whichever date occurs earlier. The rules waived by the FCC provide that ship radar stations be operated by persons licensed by the Commission in the ship service.

Radio and Television News Abroad

A British television equipment manufacturer, Pye, Ltd., has demonstrated its television transmitting equipment to FCC Commissioners and members of the staff, television representatives, members of the press, and others who attended the FCC demonstration of color and monochrome television this week. The British equipment was demonstrated November 21–22 at the Carlyle Hotel in Washington, D. C. Similar demonstrations are planned by Pye in Chicago and New York. The British concern has publicly announced its hopes to do a $5 million-a-year business in selling television transmitting equipment to U. S. broadcasters. . . . Facilities for the study of television are to be provided in the new Radio Engineering Laboratory of the Engineering College at Poona, India, according to information received by the U. S. Department of Commerce. Radio receiving tubes are produced in Spain by two companies, with total production capacity of about 2,800,000 tubes annually, according to information from the U. S. Embassy at Madrid. . . . Production of radio receivers in Austria during January-June, 1949, totalled 51,486 units, according to information received by the U. S. Department of Commerce. Production in 1948 was 96,437 sets; 21,246 in 1947; 6,175 in 1946; and 127,472 in 1937. An estimated 8,000 to 10,000 radio receivers are in use in Saudi Arabia. No import statistics are available but the U. S. and United Kingdom are believed to be the principal sources of supply, according to information received by the Department of Commerce. Several British film concerns have expressed interest in producing television films for the American market, according to information from the U. S. Embassy at London. The Devon Motion Film Studios have completed a 26-minute TV film short and sent it to this country. Other concerns have contacted the Embassy seeking information on the U. S. market. The first television broadcasting station in Latin America will begin operation in approximately four months, and will open up the initial major foreign market for U. S.-made television receivers, according to information released to RMA yesterday by the Consumers Merchandise Branch of the Office of International Trade. The U. S. Embassy at Rio de Janeiro has advised Donald S. Parris, of the OIT, that Brazilian authorities at present are "deliberating conditions under which licenses will be granted" for the importation of TV sets—"whether wholly assembled or partially assembled sets." The Embassy said there are reports that permits will be issued for large-screen sets suitable for schools, bars and public squares, and inexpensive home sets.

[1] The data on which these NOTES are based were selected, by permission, from *Industry Reports*, issues of November 18, November 25, December 2, and December 9, published by the Radio Manufacturers Association, whose helpful attitude is gladly acknowledged.

IRE People

Knox C. Black (M'29–SM'43–F'49) has recently been appointed as Technical Adviser to the Commander, Naval Air Development Center, Johnsville, Pa. This is one of a number of special appointments in the Department of Defense made under Public Law.

Dr. Black, after serving two years as an Instructor at Harvard University, joined the scientific staff of the Boonton Research Corporation, Boonton, N. J., in 1929. In 1930, he became a member of the technical staff of the Bell Telephone Laboratories, and for the next twelve years was closely concerned with research and development work on repeater and regulating equipment on the coaxial cable carrier system which is now extensively used by the American Telephone and Telegraph Company for long-distance television or multichannel telephone circuits. Beginning on a part-time basis in 1941, and later full time, he was concerned with war work. In 1943, he joined the Staff of Aircraft Radio Corporation of Boonton, N. J., but until the end of the war, the greater part of his time continued to be on war projects either at Aircraft Radio Corporation or with the Office of Scientific Research and Development. From 1948 to 1949, he held the position of Chief Engineer of Air Associates of Teterboro, N. J. He served as a member of Division 15 (and later, Division 13) of NDRC for the greater part of the war, and since that time has worked with the Research and Development Board of the Department of Defense.

In 1948 Dr. Black received a Presidential Citation for his work for OSRD during the war. In addition to his membership in the IRE, he is a member of the Phi Beta Kappa Society, the American Physical Society, the Acoustical Society of America, and the American Association for the Advancement of Science.

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Karl Hassel (A'19–M'23–SM'43) has been elected secretary of Zenith Radio Corporation. He is also a director and assistant vice-president of the corporation.

Mr. Hassel, who was born at Sharon, Pa., on January 25, 1896, studied for three years at the University of Pittsburgh. During the years 1912 until 1914, he owned and operated amateur radio equipment at Sharpsville. He then managed the radio station at the University of Pittsburgh.

After serving for one year in the Naval Radio Service, Mr. Hassel became sales manager of Chicago Radio Laboratory in 1919, and was engaged in experimental work on the development of radio apparatus for short-wave work.

Mr. Hassel began the manufacture of radio parts before the establishment of modern broadcasting systems. In 1921, as a partner in Chicago Radio Laboratories, he was manufacturing complete receivers under the trade name "Z-Nith." With Commander E. F. McDonald, Jr., he organized Zenith Radio Corp. in 1923.

Frank B. Jewett (F'20), outstanding figure in the scientific and engineering world, died November 18, 1949, at Overlook Hospital, Summit, N. J., as the result of an emergency operation. The recipient of many honors for his distinguished career in science, Dr. Jewett was a former vice-president of American Telephone and Telegraph Company, and was president of Bell Telephone Laboratories from its incorporation in 1925 until his retirement in 1940.

The Hoover Medal for 1949, one of the highest honors of the engineering profession, had been bestowed on Dr. Jewett prior to his death. The medal, which is granted by the American Society of Civil Engineers, the American Society of Mechanical Engineers, the American Institute of Mining and Metallurgical Engineers, and the American Institute of Electrical Engineers, will be awarded post-humously at the Winter General Meeting of the AIEE on January 30–February 3 in New York City, and a memorial to Dr. Jewett will be part of the program.

He was also awarded the 1950 medal of the Industrial Research Institute, Inc. for "outstanding accomplishment in leadership or management of industrial research which contributes broadly to the development of industry or the public welfare."

Dr. Jewett, who was born at Pasadena, Calif., on September 5, 1879, was graduated from Throop Polytechnic Institute (now the California Institute of Technology) in 1898. In 1902 he earned the Ph.D. degree from the University of Chicago. He then did advance study in electrical engineering and taught that subject and also physics as an instructor at MIT.

Prior to his assuming the presidency of Bell Telephone Laboratories, Dr. Jewett had served as assistant chief engineer, chief engineer, and then vice-president, of Western Electric Company, manufacturing and supply unit of the Bell System. In 1930 he was acclaimed as "one of the ten men the world could least afford to lose."

Dr. Jewett was honored by Harvard University when he was awarded the honorary degree of Doctor of Science on June 19, 1936, with a citation reading as follows: "Frank Baldwin Jewett of New York City, electrical engineer, president of the Bell Telephone Laboratories since 1925, the creator of a famous laboratory whence came miracles of modern telephony, an engineer who points the way for industry to follow."

In addition to the development of innumerable improvements in regular telephone equipment, Dr. Jewett's laboratory perfected the talking motion picture, the modern electrical phonograph, the method of transmitting photographs for long distance over telephone wires, the transatlantic telephone, and the high-speed cable.

President Roosevelt appointed Dr. Jewett to the Office for Emergency Management when that office was organized in 1940. In June of the same year the National Research Committee of the Office of Scientific Research and Development was created with Dr. Jewett as one of eight members. He was in charge of research in transportation, communication, and submarine warfare, directing its submarine warfare laboratories.

Then he became a member of the co-ordination and equipment division of the Signal Corps, and consultant to the Army Chief of Ordnance. In 1944 he was one of 12 civilian, Army, and Navy members of a committee to create a high command in science equal to the Army and Navy high commands.

At the request of the armed services, the following year he established the Research Board of National Security.

Dr. Jewett, who was given honorary degrees by eight other institutions in addition to Harvard, also received many medals. In 1939 he was awarded the John Fritz medal at MIT and was acclaimed "as one of that very small group who are the leaders, statesmen and nobleman of science." He was a former president of the National Academy of Science and the AIEE.

IRE People

Montague Ferry (SM'46), radio engineer, of Loran Research, U. S. Signal Corps, Winnetka, Ill., died recently. Mr. Ferry, whose varied career included numerous engineering associations and the formation of his own firm, was born on September 22, 1881, in Illinois.

He attended Meridian High High School, Pasadena, Calif., and received the Ph.B. degree in electrical engineering from Sheffield Scientific School of Yale University in 1902. He matriculated at MIT, obtaining the S.B. degree in 1903. The following year, he was associate professor in electrical engineering at MIT and gave lectures in "Theory of Electrical Waves." He also conducted laboratory tests.

From 1930 to 1941, Mr. Ferry established his own business as a consulting engineer, handling electric lighting and telephone estimates and supervising contracts in Winnetka. He also made check estimates of plans for a proposed subway for the city of Detroit, and worked on radio amplifier systems in his own research laboratory.

Mr. Ferry was president of the Engineering Service Corporation, with offices in New York, Pittsburgh, Chicago, and Los Angeles from 1910 to 1930. The firm engineered and contracted electric and power for industries, appraisals and war claims, and manufactured a tuned radio-frequency receiver and amplifier, among its activities.

It was in 1944 that he became associated with the U. S. Signal Corps at the Philadelphia Procurement District. Among his early business associations was service from 1906 to 1909 with the Empire Construction Co., and a year as an engineer in the research laboratory of Western Electric Co. in 1905.

Maxwell H. A. Lindsay (A'30) has been elevated to the position of chief engineer of the American District Telegraph Company, Inc., 155 Sixth Avenue, New York 13, N. Y. He has been associated with this Company since 1932.

Prior to his promotion, he was assistant chief engineer of the Controlled Companies of the A.D.T., and was responsible for the engineering requirements of all companies operations, both as to the design of equipment within the engineering department and its application through the various operating companies all over the United States. He supervised an engineering staff of more than 100 people covering the range of protection services.

From 1935 until 1946 he was engineering supervisor in charge of development and application of burglar alarm equipment and intrusion detection systems, supervised a group of engineers, and engaged in various phases of original development and application design for production and initial field trials. He was associated with Bell Telephone Laboratories from 1926 to 1932 as a member of the technical staff.

Mr. Lindsay was an instructor at MIT from 1925 until 1926. He is a member of the Acoustical Society of America, the International Municipal Signal Association, and the National Fire Protection Association.

Frank E. Spaulding, Jr. (A'35–VA'39–SM'48), supervisory engineer with the Radiomarine Corp. of America, New York, N. Y., was killed in the recent disaster in which 55 persons lost their lives in an airplane collision at Washington, D. C.

Mr. Spaulding was engaged in the design, manufacture, sales and maintenance of marine radio equipment, as well as the operation of a communication system of marine coastal radio stations as "common carrier" under a FCC license.

Born on January 18, 1910, in Connecticut, Mr. Spaulding was educated at West Haven High School and Sheffield Scientific School, and was graduated from Worcester Polytechnic Institute in 1933 with the B.S. degree in electrical engineering. From 1938 until 1945 he was a design engineer, working on the development of marine radio equipment, including receivers, direction finders, automatic alarm units, and radio telephone sets.

He was with General Electric Company at Bridgeport, Conn., from 1933 until 1938 as a test engineer in the radio receiver engineering department, conducting tests and measurements on receivers and component parts and circuit work.

L. J. N. du Treil (A'14–M'26–SM'43), a radio engineer of the Federal Communications Commission and its predecessors for the past 30 years, retired from government **service on October 31, 1949. He has formed** the company, L. J. N. du Treil and Associates, 204 Homedale Ave., New Orleans, La., and will engage in the practice of consulting radio engineering. Mr. du Treil will also establish a frequency measuring service.

Landon C. Herndon (M'36–SM'43), Assistant Chief, Field Engineering and Monitoring Division, Federal Communications Commission, died recently. Mr. Herndon was born on August 24, 1897, at **Baltimore, Md. After his graduation from high school, he attended business** college and studied electrical engineering in a special course for three years.

Mr. Herndon served in World War I with the U. S. Navy from February, 1917, until October, 1919. He was assigned to special service in sealing ship, commercial, and private radio stations; copying secret code signals from foreign and American long-wave stations for deciphering, using apparatus he had designed and built himself. He was also an instructor for Navy personnel radio operators. During active duty at sea he was chief electrician.

From 1920 until 1921, Mr. Herndon was instructor at the Norfolk Radio School, Norfolk, Va. Then he became a radio inspector with the title "assistant superintendent" of the Independent Wireless Telegraph Co., Inc., Norfolk, Va. In 1921 he became a U. S. Radio Inspector, Department of Commerce, Bureau of Navigation.

George Tompkins (A'45), production manager of Harvey Radio Laboratories, Inc., Cambridge, Mass., died recently. He was born on July 1, 1900, in New York, N. Y.

Mr. Tompkins joined Harvey Radio Laboratories in 1942 as a project engineer and contact man between the company and engineers of the armed forces and other customers interested in radio, radar, and other electronic devices.

Mr. Tompkins, who became a radio amateur in 1909 and obtained his first operator's license when he was 15 years old, commenced his business career in 1919 when he joined the test department of New York Edison Co., assisting in tests of all kinds of materials and equipment. He also did experimental work on carrier current. From 1926 until 1929 he was field technician with Columbia Phonograph Co. and the RCA Victor Co., in Camden, N. J., as supervisor of the quality division. From 1930 until 1942 Mr. Tompkins conducted his own radio service business.

Books

Modern Operational Calculus by N. W. McLachlan

Published (1948) by Macmillan and Co., Limited, St. Martin's Street, London. 211 pages+4-page index+viii pages. 24 figures. $8\frac{1}{2}\times5\frac{1}{2}$. $5.00.

This book is an excellent compendium of methods and processes for the evaluation of infinite and improper integrals. As such, it is more of a handbook on integrating procedures than a text on operational analysis. The approach is almost entirely from the real variable point of view, and the author seems to go out of his way to avoid the use of complex function theory. Within this restricted scope, however, the book does have excellent virtues. Important convergence theorems, often glossed over in engineering texts on the Laplace Transform, are here clearly presented. In a similar vein, various mathematical processes, whose validity is too frequently taken for granted, are rigorously examined. These include differentiation under the integral sign, inversion of the order of integration of multiple integrals, and differentiation and integration of infinite series.

With necessary and/or sufficient conditions established for the validity of these fundamental operations, the author proceeds to employ them in the evaluation of a wide variety of infinite integrals of which the Laplace Transform is a specific case. Dr. McLachlan also utilizes the Laplace Transform with various limiting processes as an intermediate device for evaluating and approximating many integrals. Integrands involving Bessel Functions, Error Functions, Gamma Functions, Sine and Cosine Integral Functions, and many others are so treated. The author also uses unconventional applications of the convolution integral and the infinite integral theorem to evaluate infinite integrals. This book is in fact a rich storehouse of "ingenious devices."

As mentioned earlier, the book suffers from a narrowness of scope. Thus, only incomplete consideration is given to the solution of simultaneous linear ordinary differential equations with constant coefficients, and also to linear partial differential equations. The method of solution stops short, since contour integration, and hence the systematic treatment of inverse Laplace Transforms, is avoided. Results are obtained essentially by Carson's early approach. The Laplace Transform of the solution is established and then the unknown time function is evaluated by recourse to a table.

Contour integration is employed briefly in the solution of problems arising from discontinuous periodic applied forces; e.g. sawtooths, square waves, repeated impulses, etc., and the response in these problems is obtained as a Fourier Series. Typifying the philosophy of the book, the author is more interested in this approach as an ingenious device for obtaining Fourier series of special periodic functions than as a method for solving the differential equations of a class of physical problems. The fact that closed form solutions for a typical cycle of the steady state may be easily obtained is not mentioned.

The circumscribed nature of the book is further indicated by the complete omission of the spectral concept in discussing time functions and their transforms. In this regard no mention whatever is made of the Fourier Integral, nor is reference given to the convenient table of coefficient pairs of Campbell and Foster.

Some criticism should be offered on the organization of the volume. For one thing, the footnotes are much too abundant. Further, it is generally impossible to read a section without encountering numerous references to appendices, to other sections which have appeared earlier or are yet to come, or to footnotes and figures scattered throughout the book. The ordering of material is somewhat haphazard, and the index does not seem to be as complete as the handbook character of the book warrants.

Despite the foregoing objections, this work must still be recommended. It contains a host of information on the evaluation of Laplace Transforms and other infinite integrals; and it possesses the merit, rare in engineering texts, of giving consideration to the validity of mathematical steps. It should prove a valuable reference work for engineers and physicists who employ transform analysis.

HERBERT J. CARLIN
Microwave Research Institute
Polytechnic Institute of Brooklyn
New York, N. Y.

Electric and Magnetic Fields by Stephen S. Attwood

Published (1949) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 44 pages+11-page index+xi pages+20-page appendix. 245 figures. 9×6. $5.50.

This is the third edition, with no extensive revision, of an unusual book, first written for a pedagogical purpose, which has found usage as a reference for professional engineers who at times encounter the aggravating problem of estimating a field distribution for a disarmingly simple geometry.

As a text, it is intended that the material supplement the separate introductory courses in electrical engineering by pointing out the pervasive role of the field concept in present-day analysis and design. The book seems generally well-suited to this purpose, proceeding from the more easily understood electrostatic field through the magnetic fields of currents to ferromagnetism, with a condensed treatment in the last chapter of electromagnetic fields in space and time. Emphasis is placed on return to principles, to formulate and solve typical problems. Rationalized mks units are used. Some elucidation of atomic spectra and the Bohr magnetron of dubious value is included which could better be assimilated elsewhere. There are some omissions from a table of the elementary particles in which the hydrogen atom is included.

As a reference, the book derives its chief value to practicing radio engineers from the very numerous, clear and accurate maps of electric and magnetic fields for various typical configurations, and perhaps from the extensive data on the behavior of ferromagnetic materials. The static field plots are particularly good, and will quickly recall half-forgotten details concerning potential flux in slots and around wires and corners to nonspecialists.

Adherence to practicality is the apparent reason for covering electrostatic fields in the first third of the book before mentioning Laplace's equation, and then only incidentally. However, this also leads to neglecting the beautiful and convincing demonstrations possible with the rubber membrane and the electrolytic trough, both very simple, practical, and widely used devices which circumvent trial-and-error mapping in many cases, including some three-dimensional situations. There is, consequently, no discussion of electron optics. A very short chapter treats thermionic emission and space-current flow.

Although no extensive revision has been made over the previous edition, the unique value of the book to a radio engineer continues to lie in the excellence and abundance of the field plots with their associated descriptive material.

WELLESLEY DODDS
RCA Laboratories Division
Princeton, N. J.

The A.R.R.L. Antenna Book

Published (1949) by the American Radio Relay League, West Hartford, Conn. 265 pages+5-page index. 408 figures. $6\frac{1}{2}\times9\frac{1}{2}$. $1.00.

Amateurs, experimenters, and practical radio men will find this book replete with useful information on antennas intended chiefly for amateur applications. The book is a thoroughly revised version of the previous edition.

The first five chapters deal with some of the more basic aspects of wave propagation, transmission lines, and antennas. The next five chapters describe antennas for the various amateur bands from three-quarters of a meter to 160 meters wavelength. Two more chapters are devoted to the mechanical construction of fixed and rotary arrays. No information is given on microwave antennas, such as horns or parabolic reflector types, nor is any included on waveguides.

The book has many excellent pattern diagrams and design charts, and has a considerable number of simple, easily applied formulas. References in the text and a 1-page bibliography are an aid to the advanced experimenter interested in obtaining further information.

JOHN D. KRAUS
Ohio State University
Columbus 10, Ohio

1950 IRE National Convention Program

HOTEL COMMODORE and GRAND CENTRAL PALACE—MARCH 6-9

PROGRAM

Monday, March 6, 1950

9:00 A.M.–7:00 P.M.—Registration at Hotel Commodore.

9:30 A.M.–9:00 P.M.—Registration at Grand Central Palace.

9:30 A.M.—9:00 P.M.—Radio Engineering Show, Grand Central Palace.

10:30 A.M.–12:00 M.—Annual Meeting. Principal Address by Dr. Ralph Bown, Bell Telephone Laboratories, Murray Hill, N. J., Grand Ballroom, Hotel Commodore.

2:30 P.M.–5:00 P.M.—Symposium: "Industrial Design." Symposium: "Nuclear Science and the Radio Engineer." "Applications of Semi-Conductors." "Systems I—Communication Theory." "Quality Control."

6:30–8:30 P.M.—"Get-Together" Cocktail Party, Grand Ballroom, Hotel Commodore.

Tuesday, March 7, 1950

9:00 A.M.–7:00 P.M.—Registration at Hotel Commodore.

9:30 A.M.–9:00 P.M.—Registration at Grand Central Palace.

9:30 A.M.–9:00 P.M.—Radio Engineering Show, Grand Central Palace.

10:00 A.M.–12:30 P.M.—Symposium: "Engineering for Quality in Television." Symposium: "Network Synthesis in the Time Domain." "Antennas I." "Systems II—Transmission Systems and Relays." "Measurements."

12:45 P.M.—President's Luncheon, honoring President-Elect Guy, Grand Ballroom, Hotel Commodore.

2:30 P.M.–5:00 P.M.—"Television I—Transmission Systems." "Passive Circuits—Filter Circuits and Variable Networks." "Antennas II." "Systems III—Modulation Systems and Bandwidth Requirements." "Industrial Instruments."

8:00 P.M.–10:30 P.M.—Symposium: "Television."

Wednesday, March 8, 1950

9:00 A.M.–7:00 P.M.—Registration at Hotel Commodore.

9:30 A.M.–6:00 P.M.—Registration at Grand Central Palace.

9:30 A.M.–6:00 P.M.—Radio Engineering Show, Grand Central Palace.

10:00 A.M.–12:30 P.M.—"Television II—UHF and Color TV." "Computers I—Digital Computers." "Electron Tubes I—Theory and Design." "Active Circuits I—Amplifiers." Symposium: "Basic Circuit Elements."

2:30 P.M.–5:00 P.M.—"Television III—Receivers." "Active Circuits II—General." "Electron Tubes II—Theory and Design." "Computers II—Information Analysis and Computing." "Transmission and Antennas."

6:45 P.M.—Annual IRE Banquet (dress optional), Grand Ballroom, Hotel Commodore. Awarding of the Medal of Honor, the Morris Liebmann Memorial Prize, the Browder J. Thompson Memorial Prize, the Harry Diamond Award, the Editor's Award, and Fellow Awards.

Thursday, March 9, 1950

9:00 A.M.–2:30 P.M.—Registration at Hotel Commodore.

9:30 A.M.–6:00 P.M.—Registration at Grand Central Palace.

9:00 A.M.–6:00 P.M.—Radio Engineering Show, Grand Central Palace.

10:00 A.M.–12:30 P.M.—"Audio-Transducer Design." "Electronics in Medicine." "Propagation I—Propagation at Ionospheric Frequencies." "Electron Tubes III—Power Tubes." "Navigation Aids."

2:30 P.M.–5:00 P.M.—Symposium: "Sound Recording." "Propagation II—Impact of Propagation on Operation of Systems." "Electron Tubes IV—Materials and Techniques." "Components." "Oscillators."

Committee Meetings

The following IRE committee meetings have been scheduled at the Hotel Commodore during the convention:

Professional Groups Committee (including Group Chairmen)
Chairman, W. R. G. Baker
Tuesday, March 7, 8:45 A.M.–10:00 A.M., Parlor B

Sections Committee (annual meeting)
Chairman, J. F. Jordan
Tuesday, March 7, 5:00 P.M.–7:00 P.M., West Ballroom

Professional Group Chairmen and Group Members
Chairman, W. R. G. Baker
Wednesday, March 8, 8:45 A.M.–10:00 A.M., West Ballroom

Professional Groups Sponsored by the Measurements and Instrumentation Committee
Chairman, Ernst Weber
Thursday, March 9, 8:45 A.M.–10:00 A.M., West Ballroom

Audio Professional Group (annual meeting)
Chairman, L. L. Beranek
Thursday, March 9, 9:00 A.M.–10:00 A.M., Grand Ballroom

Standards Committee
Chairman, J. G. Brainerd
Thursday, March 9, 1:00 P.M.–2:30 P.M., Parlor B

All chairmen and vice-chairmen of technical committees are requested to be present at the Standards Committee meeting in order to present all suggestions for technical committee operations to the Standards Coordinator and the Chairman of the Standards Committee. The Standards Co-ordinator is looking forward to greeting the chairmen of all Technical Committees personally. This meeting will also provide the first opportunity for a get-together with the chairmen of the several new technical committees created during the past year.

Annual Meeting

MONDAY, MARCH 6

10:30 A.M.

This opening meeting of the convention is for the entire membership. The meeting will feature as the principal speaker Dr. Ralph Bown, Bell Telephone Laboratories, Murray Hill, N. J.

Women's Program

Monday, March 6, 1950

12:30 P.M.–1:30 P.M.—TV Program, DuMont Studios, Madison Avenue and 53 St. (Limited to first 49 reservations).

P.M.— "Get Acquainted" at Ladies' Headquarters, Hotel Commodore.

Tuesday, March 7, 1950

9:30 A.M.—"Behind The Scenes" Tour – Museum of Natural History, Central Park West and 78 St.—Luncheon at the Museum. (Transportation included).

3:00 P.M.–5:00 P.M.—Tea at IRE headquarters, 1 East 79 St., New York.

Wednesday, March 8, 1950

9:30 A.M.—Tour of Good Housekeeping Institute, Eighth Avenue and 57 St. Luncheon at Castleholm. (Transportation included).

2:30 P.M.—Matinee (choice): "Miss Liberty," "Where's Charley," or "Death of a Salesman."

Thursday, March 9, 1950

9:00 A.M.—Tour of Cunard Liner, "Queen Elizabeth," if in port; alternate—trip to Metropolitan Museum to view Hapsburgh collection. (Transportation included).

Note

No papers are available in preprint or reprint form nor is there any assurance that any of them will be published in the PROCEEDING OF THE I.R.E., although it is hoped that many of them will appear in these pages in subsequent issues.

SUMMARIES OF TECHNICAL PAPERS

SYMPOSIUM
Industrial Design

1. EVOLUTION AND GROWTH OF INDUSTRIAL DESIGNING

JOHN VASSOS
(Designer, New York, N. Y.)

This paper will cover the advantages of industrial design in product planning, and the need for better looking equipment; designing for present day trends of pleasing appearance in addition to good functional engineering; a short history of industrial design and how it has progressed over the years from its initial acceptance during the 20's, its growth to a recognized industrial need during the 30's, and then on into the post-war era and its importance in today's product merchandising; examples of the evolution of design trends as they have progressed through the automotive industry, modern railroading, and commercial aviation; examples of the great strides in the refrigeration industry, with comparisons such as the monitor top vs. present clean lined design, and the mantle-type radio of the early 30's vs. the very up to date and modern conceptions of television.

2. PROCEDURE IN INDUSTRIAL DESIGNING

CLYDE PETERSON
(Westinghouse Electric Corporation, Sunbury, Pa.)

This paper will be a discussion of the development of a console television cabinet through the various considerations and steps of its progress from inception to production; field surveys, dealer, and consumer interviews and how they are used to determine the most acceptable types and styles; when correlated and after broad discussions how rough sketches and designs are prepared which are developed into two or three of the most favorable candidates; the decision of the number of samples to be made; full size drawings prepared of the selected designs, and complete samples built by cabinet vendors after a thorough understanding has been established with them as to construction, color, and finish; the presentation to Sales and Management of the final determination of the model to be used. The manner in which the designer and the engineer work through these stages, and the presentation of final drawings, and the final production release will be discussed.

3. COST REDUCTION POSSIBILITIES IN INDUSTRIAL DESIGN

W. B. DONNELLY
(General Electric Co., Syracuse, N. Y.)

A good design costs no more, and oftentimes it is possible that it may be less expensive, than poorly designed equipment. One of the prime requisites in obtaining the greatest value for any given cost, is adequate design time in the planning schedule so that the greatest advantage can be obtained from the studied use of the best materials and techniques to be utilized. Hurried designs often result in high costs, because of the lack of proper consideration of manufacturing techniques. Complete exploration of lower costs and more attractive materials, added to a close liaison and complete understanding with engineering staffs, will make for good design at lower cost.

4. SALES ATTITUDE TOWARDS INDUSTRIAL DESIGN

E. P. TOAL
(North American Philips Co., Inc., New York, N. Y.)

In today's selling, attractive designing is one of the prime factors in sales acceptability. Those persons engaged in sales, from the distributor to the retail salesman, are 100 per cent sold on the value of good design. It is one of the most important factors in consumer preference, all other considerations being equal. The relationship between industrial design and sales planning, and the industrial design activities of the major companies will be discussed, along with the growth of independent design organizations, which today point to the advantages and the preference of sales acceptance of good design.

SYMPOSIUM
Nuclear Science and the Radio Engineer

5. NEWS OF THE NUCLEUS

URNER LIDDEL
(Office of Naval Research, Washington, D. C.)

Knowledge of atomic nuclei is expanding rapidly. Every radio engineer is well acquainted with the antics of electrons. New particles appearing on the horizon will, undoubtedly, be of importance to him in the next several years. Nuclear engineering will require a close tie with electron engineering.

Concepts of the basic particles of matter have undergone a complete cycle in the last 2,000 years. The trend is always toward greater complexity until some order can be found which simplifies the ideas involved. The evolution of some of these ideas will be discussed. The characteristics of basic particles will be delineated, and our present knowledge concerning them described.

6. PARTICLE ACCELERATORS

M. STANLEY LIVINGSTON
(Massachusetts Institute of Technology, Cambridge, Mass.)

The particle accelerators developed before 1945 were based on relatively simple physical and engineering concepts; these were the standard cyclotron, the betatron, and the several machines for generating direct high voltages of which the best known is the Van de Graaff generator. Modern particle accelerators capable of developing much higher particle energies are based on the principle of synchronous acceleration through hundreds of thousands of small steps. An essential part of this principle is the "Phase stability" resulting from the proper combination of electric and magnetic fields to achieve focussing both in the spatial coordinates of the orbit and in particle energy. The synchro-cyclotron, electron synchrotron, and linear accelerator are all designed to utilize such phase stability. A common characteristic of particle motion in these machines is the "bunching" in phase of crossing the accelerating gap, and the phase oscillations in energy and location which maintain the bunching through an extremely large number of accelerations. The practical energy limits of these machines occur for different reasons but at approximately the same limit of about 500 million electron volts. The proton synchrotron is the one synchronous accelerator capable of achieving energies in the 1- to 10-billion volt range. These several accelerators will be described and the physical principles of the particle motions discussed.

7. RADIO-FREQUENCY PROBLEMS ASSOCIATED WITH PARTICLE ACCELERATORS

J. P. BLEWETT
(Brookhaven National Laboratory, Upton, L. I., N. Y.)

In the previous abstract, Livingston has pointed out that most of the modern high energy particle accelerators depend on synchronous acceleration in radio-frequency fields. The cyclotron, the linear accelerator, and the synchrotron all belong to this class. In the ordinary cyclotron, the linear accelerator and the electron synchrotron, the frequency is held constant, but in the frequency-modulated cyclotron and the proton synchrotron, the frequency must change during the accelerating cycle, sometimes by factors of ten to one or more. The generation and frequency control of the rf signal, and its application to the particles under acceleration all present novel problems of considerable difficulty. In the FM cyclotron, the frequency range is relatively narrow and mechanical tuning is possible, but in the proton synchrotron, precise timing requirements make electrical frequency control essential. This paper will be concerned with the new techniques which have been developed in the solution of these problems.

8. DETECTION OF NUCLEAR RADIATIONS

JOHN R. DUNNING
(Columbia University, New York, N. Y.)

The basic characteristics will be reviewed of the present available devices for detecting nuclear radiations, including heavy and light charged particles, neutrons, mesons, and quanta. The limitations and future possibilities of ionization chamber systems, linear amplifiers, geiger and proportional counters, fluorescent and conductivity counters, and general multiplier techniques will be discussed. The role of essentially electronic devices as contrasted with nonelectronic devices such as photographic and cloud chamber methods will be considered.

Special emphasis will be placed on the problems of detecting radiations with energies over the range from zero to 1 B.e.v. and higher, and the problems introduced by pulsed operations of radiation sources.

Applications of Semi-Conductors

9. TRANSISTOR TRIGGER CIRCUITS

HERBERT J. REICH AND PETER M. SCHULTEISS

(Yale University, New Haven, Conn.)

Following an introductory discussion of the characteristics of transistors in the light of their application to trigger circuits, the authors will discuss several types of trigger circuits employing transistors. Some of these circuits use transistors only, some use transistors and crystal diodes, and some use transistors and vacuum tubes. Oscillograms, which will be shown, indicate that the switching time is of the order of 0.1 microsecond.

10. GLASS-SEALED GERMANIUM DIODES

S. F. AMICO

(Sylvania Electric Products Inc., Boston, Mass.)

Improvements in germanium diodes have been constantly directed towards reduction in size, improvement in stability, and resistance to extremes of temperature and humidity.

The germanium diode has now been hermetically sealed in a tiny glass envelope. Incidental with this process were many mechanical and electrical problems, not the least of which was a method of individual inspection of diodes to eliminate those having imperfect hermetic seals. This glass-sealed diode which needs no wax filling has proved highly successful in meeting the requirements established as is attested by test results.

11. HIGH-TEMPERATURE CHARACTERISTICS OF GERMANIUM DIODES

LOWELL S. PELFREY

(Raytheon Manufacturing Co., Newton, Mass.)

This paper will present data on reverse characteristics of germanium crystal diodes from 0 to 100,000 cycles, and from 25° to 100° C.

The data will be in the form of reproduction of oscillograph traces of the diode characteristic over this frequency range. Measurements of rectification efficiency over these ranges of temperature and frequency will be made and correlated with the reverse resistance data. The frequency range was selected so as to be as wide as possible, and still keep capacitance effects at a minimum, and the temperature range is one commonly encountered in actual service.

12. MATHEMATICAL THEORY AND APPLICATIONS OF SILICON CRYSTALS FOR MIXING AND HARMONIC GENERATION AT MICROWAVE FREQUENCIES

P. D. STRUM, J. W. KEARNEY, AND J. C. GREENE

(Airborne Instruments Laboratory, Inc., Mineola, L. I., N. Y.)

This paper presents a mathematical theory, based solely upon the static voltage-current characteristics of silicon crystals, of both fundamental and harmonic mixing and harmonic generation. The paper also describes experimental results, obtained at both audio and microwave frequencies, which verify the theory. The conversion loss and harmonic generation efficiency which can be obtained from a crystal are determined by the average value of the parameter, $\frac{d\,(\log i)}{d\,(\log e)}$ for the particular operating conditions. Design criteria for optimizing this parameter are given for the various modes of operation. The theory also allows the computation of optimum impedance and noise temperature conditions.

13. CONSIDERATIONS IN LOW NOISE FIGURE MICROWAVE RECEIVER DESIGN

M. T. LEBENBAUM AND P. D. STRUM

(Airborne Instruments Laboratory, Inc., Mineola, L. I., N. Y.)

This paper discusses several factors which assume increasing importance as the noise figure of microwave receivers approaches more closely that of the theoretically perfect receiver.

It is shown that the direct frequency dependence of the minimum attainable if noise figure and the inverse frequency dependence of the excess noise of a crystal converter establish an optimum frequency for the if amplifier for minimum over-all noise figure.

Methods are suggested for reducing the crystal excess noise by about 50 per cent. The effect of finite coil Q's as a limiting factor in if input circuits is discussed. The effects of image and other spurious responses on noise figure and its measurement are discussed in detail.

Systems I

Communication Theory

14. REPRESENTATIONS OF SPEECH SOUNDS AND SOME OF THEIR STATISTICAL PROPERTIES

SZE-HOU CHANG, GEORGE E. PIHL, AND MARTIN W. ESSIGMANN

(Northeastern University, Boston, Mass.)

Spectrographic analysis, autocorrelation, and infinite clipping are considered as methods of transforming and analyzing speech sounds with the object of obtaining simple representations without excessive loss of intelligibility. Although not mathematically equivalent in the exact sense, these methods when properly approximated are shown to provide parameters of statistical nature that are simply related. It is conjectured that the parameters describe some essential elements of speech sounds, and are statistically invariant. Experimental techniques for performing the analyses and checking some of the results are described.

15. SPEECH TRANSMISSION THROUGH RESTRICTED BANDWIDTH CHANNELS

M. J. DITORO, W. GRAHAM, AND S. SCHREINER

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

It has been recognized for some time that the present bandwidth of about 3,000 cps used for speech transmission is much larger than required to transmit the speech of one person with satisfactory articulation and speaker recognition. This is because the speech of one person does not fill up all the 3,000 cps bandwidth allotted to it all of the time. For example, during pauses between words or sentences, the bandwidth is idle.

Another reason is the redundancy arising because of the quasi-periodic nature of the voiced sounds. The Fourier transforms of the latter indicate a partially-utilized spectrum characterized by a number of narrow band clusters (or sidebands) around harmonics of the fundamental frequency or pitch of the sound. An equivalent electrical network is derived for the acoustic cavities of the human speech system, from larynx to mouth, which indicates why these redundancies exist and also indicates, by its natural modes, the seat of the information content in speech.

Various electronic methods are considered by which these redundancies may be exploited to reduce the bandwidth required for the transmission of the speech of one person. Consideration is given to methods which give moderate compression of the bandwidth with reasonable equipment size, and also, at the other extreme, to those methods which result in a considerable reduction in bandwidth but with consequent increase of equipment size.

16. APPLICATION OF COMMUNICATION THEORY TO PERIODIC RADIO SYSTEMS

M. LEIFER AND N. MARCHAND

(Sylvania Electric Products Inc., Bayside, L. I., N. Y.)

The statistical theory of Wiener and Shannon is used as a basis for the choice of the values of the fundamental parameters of bandwidth and power for periodic radio systems. The term "periodic system" is here taken to mean any system which employs a repetitive type of transmission, such as radar. The transmitted signals do not convey information since these signals are known, but they do establish a basis for generation of information through space modulation or absorption modulation between transmission and reception. The problems of detection of the signals when obscured by interference and noise are shown to be related to corresponding problems in such diverse fields as spectroscopy and recognition. Correlation of the received signals with standard signals is employed in the solution of these problems, and this method of analysis leads naturally to emphasis upon the spectral distribution of such signals.

17. SOME ASPECTS OF DATA TRANSMISSION OVER NARROW-BAND COMMUNICATION CIRCUITS

MILLARD M. BRENNER
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

This paper outlines some problems encountered in transmitting signals representing precise time-variable position data over noisy, narrow-band transmission facilities. Since instantaneous input-output error must be kept very small, precision in both message value and message timing must be maintained. Available carrier-to-noise levels require an expansion of the original data message bandwidth in order that the instantaneous message value may be transmitted with high precision. Transmission delay due to finite bandwidth has fixed and variable components. Dynamic errors due to delay can be compensated for to an extent, some error being due to the uncertainty component of the delay, which is inversely related to bandwidth and carrier-to-noise ratio.

18. THE STATISTICAL PROPERTIES OF NOISE APPLIED TO RADAR RANGE PERFORMANCE

S. M. KAPLAN AND R. W. MCFALL
(General Electric Co., Schenectady, N. Y.)

This paper deals primarily with the performance of radars having intensity modulated displays, i.e. displays which convert signal amplitude to cathode ray tube spot intensity. The advent of high speed scanners, which yield few pulses per target, using intensity modulated displays has necessitated an analytical approach, as empirical methods have proven inadequate. Statistical concepts involved in the effect of receiver noise on the display are analyzed. This analysis yields several graphs from which the radar detection range can be calculated for any desired detection probability, and from which computations of the probability of false targets echoes can be made.

Quality Control

19. ACCELERATED LIFE TESTING OF VACUUM TUBES

JEROME ROTHSTEIN
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

Industry and the Armed Services need accelerated tube testing procedures for economic and military reasons, respectively. Tube failure occurs in many forms whose relative frequencies depend on manufacturing and operational variables. A survey of failure data covering 90 per cent of the industry has been made and analyzed, and attempts at accelerated testing evaluated. The philosophy of government sponsored work will be outlined. It is to analyze important failure types, devise conditions accelerating them individually, correlate accelerated with normal failure rate for each type, and from the former synthesize information equivalent to conventional life data. Difficulties and alternative plans will be discussed.

20. STATISTICAL EVALUATION OF LIFE EXPECTANCY OF VACUUM TUBES DESIGNED FOR LONG-LIFE OPERATION

ELEANOR M. MCELWEE
(Sylvania Electric Products Inc., Kew Gardens, L. I., N. Y.)

Life-test data on subminiature vacuum tubes designed for 5,000 hours are analyzed statistically and an equation is derived for the curve of life survival percentages. Correlation of individual types to the general curve is found to be extremely high. Controls are determined for normal 500-hour life tests which assure rated long-life quality and are presented as a method for evaluating life expectancy before completion of long-life tests. Life test samples of lots of tubes released by this 500-hour plan were continued in operation for 5,000 hours, and the results are shown to be satisfactory.

21. APPLICATION OF STATISTICS TO ACCEPTANCE SPECIFICATIONS

BARTON KOSLOW
(Allen B. DuMont Laboratories, Inc., Clifton, N. J.)

Specification by a design engineer without production knowledge often results in uneconomical specification. Using the experience from a pilot or production run has greater merit. A simple statistical analysis of this information has the advantage of greater efficacy and accuracy of prediction from the same data.

The use of control charts and the fiducial probability of a mean is illustrated. Further classification of basic assumptions and procedures is then made.

How the statistical method overcomes the difficulties of the usual types of specification is discussed. In conclusion, a point-by-point comparison of the different specification methods is achieved.

22. STATISTICAL METHODS IN RESEARCH AND DEVELOPMENT

L. LUTZKER
(Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

The statistical method, called "Analysis of Variance," is used when one wishes to assess the relative importance of various factors contributing toward variation in a desired result. An experiment, involving variation in the factors under study, is designed so that it will yield data in systematic groups. The source variance among groups is then compared with the error variance and the significance of the result determined. By this method of analysis, those factors exerting significant influences on the final data are established. In addition, the existence and extent of interactions among the factors examined are ascertained.

23. STATISTICAL ENGINEERING OF TOLERANCES

EUGENE D. GODDESS
(Sylvania Electric Products Inc., Boston, Mass.)

Inherent in mass production is the implication of selection of stochastic variables, i.e., those chosen by chance. Hence the determination of over-all tolerances is governed by those aspects of compound probability which obtain in these situations. Probability theory will be discussed as it affects the theoretical basis for addition of tolerances. Reference will be made to the addition of tolerances when the universe is normal, either complete range or truncated. Design examples of radio components and circuits will be given.

SYMPOSIUM
Engineering for Quality in Television

24. TOP MANAGEMENT EVALUATES QUALITY IN TERMS OF SOUND ENGINEERING

ALLEN B. DUMONT
(Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

Although quality of the product depends upon many things, one of the most important of these, perhaps, is quality of design. Therefore, while covering several broad phases of Engineering for Quality, Dr. DuMont will discuss the part played by the technical man, not only as applied to the design function, but also to the many associated functions as well.

25. STATISTICS—A NEW TOOL FOR THE PLANNING AND ANALYSIS OF LABORATORY EXPERIMENTS

ENOCH B. FERRELL
(Bell Telephone Laboratories, Inc., New York, N. Y.)

The statistical methods that have been developed for use in quality control are a powerful tool in the interpretation of laboratory experiments where only a small amount of data are available. An understanding of these methods also permits more logical planning of experiments and improves what we might call "the efficiency of experimentation." One of the simplest and most broadly useful of these tools is the control chart. It is easy to understand and use and in many cases can take the place of more laborious and complicated methods of analysis.

26. SPECIFICATION FOR QUALITY OF THE VISUAL OUTPUT OF PICTURE TUBE SCREENS

A. E. MARTIN
(Sylvania Electric Products Inc., Bayside, L. I., N. Y.)

Among the several significant aspects of quality control as applied to television receiver manufacture are those which specify the visual output from picture tube screens. They are: chromaticity, luminance and contrast. A review is given of the problems incidental to the measurement of each named aspect. There follows a discussion of the present status of such measurements in the television industry, with emphasis upon the availability of adequate standardization. The importance of this is made clear by the fact that the correct interpretation of data by the I.C.I. system depends upon the maintenance and use of satisfactory standards. The conclusion of these remarks embodies a delineation of further work needed for the firm establishment of quality specifications.

27. THE IMPORTANCE OF PRACTICAL DESIGN AND SPECIFICATIONS FOR EFFECTIVE PRODUCTION AND HIGH QUALITY

J. MANUELE
(Westinghouse Electric Corporation, East Pittsburgh, Pa.)

An almost perfectly designed television set may be a work of art and a triumph to science. It may out-perform contemporaries, but unless it can be produced en masse and with manufacturable ease, it has very little chance for success. In his paper, Mr. Manuele will elaborate on the need for a practical design, that is, one which is compatible with present facilities for production, for adequate covering specifications, and for a system of quality control by which an item can be evaluated from raw material to finished product.

SYMPOSIUM

Network Synthesis in the Time Domain

28. A COMPARISON OF FREQUENCY AND TIME DOMAIN VIEWPOINTS IN CIRCUIT DESIGN

W. H. HUGGINS
(Air Force Research Laboratories, Cambridge, Mass.)

Many familiar concepts of frequency filtering have their counterparts in the time domain. With the recent development of pulse communication systems and information theory, the treatment of signals as actual functions of time, rather than as some equivalent spectrum of frequency components, is a matter of growing importance.

The classical relations between time and frequency formulations of network theory, will be reviewed with emphasis on methods of interpreting the characteristics of networks and signals in the frequency and time domain. Two possible design criteria formulated in the time domain will be described and the need for a more general approach which combines both frequency and time viewpoints is indicated.

29. STUDY OF TRANSIENT EFFECTS BY A NEW METHOD OF INTEGRAL APPROXIMATION

M. V. CERRILLO
(Massachusetts Institute of Technology, Cambridge, Mass.)

The characterization of the nature of a transient at different intervals of time and the correlation of wave forms with the analytical elements of the system and the excitation functions constitute one of the basic studies in the foundation of the theory of synthesis of discrete and distributed electrical and similar linear systems.

By means of certain methods of approximate integration, here described, the above goal can be fairly well attained. Pertinent theorems of transient equivalence have been discovered, which led to the concept of "generating functions." Some other important relations between the time and frequency domains can be derived, which are of great help in showing the proper approach to the solution of certain problems of network synthesis.

30. APPLICATIONS OF THE INTEGRAL APPROXIMATION METHOD OF TRANSIENT EVALUATION

W. H. KAUTZ
(Massachusetts Institute of Technology, Cambridge, Mass.)

The principal features of the method of approximate integration as applied to linear, lumped-constant networks are illustrated with a number of examples. In particular, the response of networks to a particular class of applied transient stimuli, including frequency modulated transient signals, is discussed from a general standpoint to illustrate the correlation between the time and frequency domains. Finally, an example of the synthesis of networks for prescribed transient and spectrum response is worked out to demonstrate the approximation techniques involved and resultant forms of the transient solution.

31. TRANSIENT RESPONSE OF ASSYMETRICAL CARRIER SYSTEMS

G. M. ANDERSON AND E. M. WILLIAMS
(Carnegie Institute of Technology, Pittsburgh, Pa.)

It is known that systems such as vestigial sideband television systems having asymmetrical amplitude and phase characteristics about the carrier frequency possess a certain type of nonlinearity in response even though the system is linear in the total signal. The paper describes a vector integral method of time-domain analysis, analogous to the super-position integral for linear systems, for determining transient response to arbitrary modulation. Typical experimental responses are shown and analytical means are given for placing limits on the transient response variation as a function of the signal modulation. Conditions are pointed out which give unusual singularities in the transient response.

Antennas I

32. WAVEGUIDE APPLICATIONS OF ARTIFICIAL METALLIC DIELECTRICS

W. E. KOCK
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

Artificial dielectrics consist of arrays of metallic elements whose size and spacing are small compared with the wavelength. The polarization effect of a true dielectric is duplicated through the polarization of the individual dipole elements. Originally employed as lightweight substitutes for true dielectrics in microwave lens antennas, they also find use in other microwave applications where true dielectrics would ordinarily be employed. The present paper describes the use of iterative dipole structures as variable phase shifting elements in waveguides, as wave-guiding structures both in free space and over conducting plates, and as waveguide mode converters. A longitudinal row of transverse dipoles in which the plane of dipoles is gradually rotated provides a polarization rotation device also useful for waveguide mode converters.

33. THE EFFECTS OF ANISOTROPY IN A THREE-DIMENSIONAL ARRAY OF CONDUCTING DISKS

GERALD ESTRIN
(University of Wisconsin, Madison, Wis.)

This microwave delay lens medium is shown to have both magnetic and electric anisotropy, which necessitates an analysis describing its refractive properties for obliquely incident waves.

A simple linear transformation is applied to the field equations such that the transformed system is magnetically isotropic. Classical solutions from studies in optics provide the ray velocity surfaces in that system. An inverse transformation yields the ray velocity surfaces in the original medium.

Huyghens construction is employed, for two particular arrays, to determine the direction of the refracted wave after oblique incidence.

34. A STUDY OF SINGLE-SURFACE CORRUGATED GUIDES

WALTER ROTMAN
(Air Force Research Laboratories, Cambridge, Mass.)

A series of experiments were conducted to determine the properties of single surface corrugated structures as transmission lines and as radiators. Two types of surfaces were tested: the first is a flat grooved plate fed by a waveguide, and the second a circular corrugated cylinder of "disk-on-rod" structure fed by a coaxial line. A modification of the second type results in a spirally grooved rod with similar properties.

The measured field parameters are found to be predictable from existing theory. For properly designed structures the energy is essentially bound to the corrugated surface; little radiation occurs, and the attenuation of the traveling wave is due chiefly to losses in the metal. The effect of filling the corrugations with solid dielectric is also subject to analysis.

Although attempts to use the various units as antennas by radiating from discrete points on the surface present distinct disadvantages, tapering the corrugations to match the guide impedance to that of free space has met with some success.

35. A STUDY OF THE CURRENT DISTRIBUTION ON THE HELIX

JAMES A. MARSH
(Ohio State University, Columbus, Ohio)

The current distribution on a uniform circular helix has been measured over a three-to-one frequency band which spans the axial mode of radiation. The measured distributions have been analyzed in terms of traveling waves of current which may be associated with three different transmission modes on the helix. The relative amplitude functions of these traveling waves, as well as their associated phase velocities, have been approximated. Current distributions which have been calculated by superposing three or more of these traveling waves are in good agreement with the measured data.

Work described in this paper was carried out, in part, under a contract between the Wright-Patterson Air Force Base of the Air Matériel Command and the Ohio State University Research Foundation.

36. DIFFRACTED BEAMS IN METAL LENSES

Albert E. Heins
(Carnegie Institute of Technology, Pittsburgh, Pa.)

If a plane electromagnetic wave is incident upon an infinite set of staggered, equally-spaced semi-infinite parallel planes, there will be at least one reflected wave in free space and possibly a transmitted wave in the ducts, depending on the wavelength, the spacing, and the polorization. For the *E*-plane case (discussed by Carlson and Heins for the case of a single reflected and transmitted wave) we show that diffracted beams may arise by modification of the original conditions on wavelength, angle of stagger of the parallel plates, and angle of incidence of the plane wave. The results are also presented numerically.

Systems II

Transmission Systems and Relays

37. SIGNAL CORPS HIGH-FREQUENCY RADIO COMMUNICATION RESEARCH AND DEVELOPMENT

John Hessel and H. F. Meyer
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

The role of hf radio communication in the military communication system is described. A statement is made of the technical problems requiring solution in order to improve hf communications. The integrated program of research and development now being pursued by the SCEL to accomplish these improvements is presented. The subjects covered include modulation studies, diversity investigations, stable frequency generating sources, single sideband techniques, radioteletype advances, and system integration analyses. The inter-relation of the various projects to improved communication equipment, facilities, and systems is discussed.

38. MILITARY SINGLE SIDEBAND EQUIPMENT DEVELOPMENT

R. A. Kulinyi
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

This paper gives a review of Signal Corps Engineering Laboratories' activity in single sideband (SSB) development. Characteristics are given of a typical transmitting and receiving system presently used by the military services. Experimental systems are described which employ audio phase-shift net works instead of channel filters. One type of transmitter uses sideband generators which operate at a low level, followed by linear amplifiers. Another experimental transmitter will generate the desired sideband at a high power level. Projects are described which lead to both high- and low-frequency embodiments of the high level principle. Adapters for SSB reception with double sideband receivers are considered. One, already under test, uses phase-shift networks. A second adapter developed at this time is intended to evaluate all available means of sideband selection to provide best obtainable performance. A brief description is given of a test program for comparing various SSB systems and some instrumentation problems are considered.

39. RADIO RELAY DESIGN DATA 60 TO 600 Mc

R. Guenther
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

This paper presents general design data for a complete radio relay transmission system including propagation characteristics (60 to 600 Mc) multichannel operation and various types of modulation. All parameters are determined by the noise figure of the receiver. From these figures either a level diagram may be plotted as usual in communication circuits, or the final signal to noise ratio per channel may be obtained from simple addition or subtraction of the various parameters respectively. Thus, a convenient procedure for practical use is obtained which permits a quicker and reliable design. In order to extend the figures to various types of modulation a new parameter has been defined which takes care of the efficiency of the demodulator.

40. MULTIPLEX MICROWAVE RADIO RELAY

D. D. Grieg and A. M. Levine
(Federal Telecommunication Laboratories, Inc. Nutley, N. J.)

A description is given of a 5,000-Mc 24-channel radiotelephone relay system suitable for both short- and long-haul transmission.

Multiplexing of the voice channels is accomplished on a time-division basis, together with pulse-time modulation of the channel pulses. The rf is modulated by means of frequency-shift keying with the pulses and spaces being at different carrier frequencies.

The relay system consists of terminal equipment for accomplishing the time-division multiplex and pulse-time modulation, and rf transmitters and receivers for the microwave transmission. The repeater comprises a transmitter and receiver connected back to back, and pulse frequencies rather than audio frequencies are handled.

The main features of this system include the minimization of cross stalk and distortion coupled with improved transmission characteristics.

41. CROSS TALK IN FREQUENCY- AND PHASE-MODULATED RADIO RELAYS USED IN CONJUNCTION WITH MULTICHANNEL TELEPHONY EQUIPMENT

Saul Fast
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

In transmitting the frequency division multiplexed output of a carrier telephone terminal over a radio relay, the achievement of high signal to cross talk ratios presents considerable difficulty. A general method is developed for treating cross talk arising in nonlinear circuits. This method is applied to direct and indirect phase and frequency modulations coupled circuit, and single-tuned interstages and discriminator circuits. The effect of mistuning is also treated.

Relations are obtained between cross talk to signal ratio per channel and harmonic distortion for each of these circuits. There is considerable difference in cross talk to signal ratio per channel when frequency or phase modulation is employed.

The effect of cascading distorting circuits on the cross talk is considered and methods of minimizing cross talk are presented.

Measurements

42. OSCILLOGRAPHIC PRESENTATION OF TIME DELAY AND DISTORTION IN BROAD-BAND FM SYSTEMS

A. R. Vallarino
(Federal Telecommunication Laboratories, Inc. Nutley, N. J.)

Two panoramic equipments have been developed to help in the design of broad-band FM systems over the range of 100 kc to 10,000 Mc. Maximum bandwidth is 40 Mc. These are as follows: (1) An equipment to measure the incremental time delay (phase-frequency slope) of intermediate transmission networks. (2) An equipment to measure the amplitude-frequency slope characteristics of modulator and demodulator components.

Incremental time delays of 1 millimicrosecond (10^{-9} secs) and static harmonic distortions of 70 db can be measured. In both cases the slope characteristic is presented on an osciloscope. Time delay measurements can be made from 100 kc to 5000 Mc. Distortion measurements from 10 Mc to 10,000 Mc.

43. ROCKETS RANGE INSTRUMENTATION

E. R. Toporeck and F. M. Ashbrook
(U.S. Naval Ordnance Test Station, Inyokern, Calif.)

The problem of obtaining ballistic information for the study and evaluation of the performance of rockets and guided missiles is complex. It is necessary to determine such quantities as position in space, trajectory, acceleration, roll, pitch, yaw, etc. Electronic and photographic equipment are used to obtain the necessary information on missile performance. Electronics provide such items as precise timing, electronic and magnetic sensing avices including telemetering, radar equipment, and recording techniques. The photographic equipment is used to make graphical records of functions and to record the position in space of the missile, which is correlated with the electronic for time definition, etc.

The diversity encountered in an experimental program such as is performed at NOTS requires unusual flexibility of precise instrumentation of many and varied kinds. These requirements have lead to the development of many unusual techniques in the instrumentation field.

44. NEW TEST EQUIPMENT FOR THE UHF TELEVISION BAND

John Ebert and H. A. Finke
(Polytechnic Research and Development Co., Inc., Brooklyn, N. Y.)

This paper covers the important electri-

cal and mechanical design solutions for three new pieces of test equipment that were developed for the uhf television band. All three have novel electrical solutions that combine distribution and lumped circuit techniques. The test equipments to be discussed are:

1. A sweep-frequency oscillator operating at fundamental frequency.

2. A novel direct-reading cavity-type frequency meter and marker. The drive on this instrument has been linearized to a degree that permits checking frequency differences to 0.1 Mc per second.

3. A calibrated variable noise source using a coaxial noise diode.

45. DIRECT READING PHASEMETER

L. H. O'Neill and J. L. West
(Columbia University, New York, N. Y.)

A device for measuring the phase difference between two sinusoidal voltages in the frequency range 7 cps to 100 kc is described. The method employs the summation or subtraction of the two voltages. Accurate summation and subtraction is performed by the use of feedback amplifiers. The phase difference may be read directly on a vacuum-tube voltmeter. The precision of measurement may be made as high as is desired through the use of multistage feedback amplifiers. However, precisions of the order of 1/2° are obtained by the use of relatively simple circuitry.

46. MEASURING PROCEDURE FOR RADIOTELETYPE CONVERTERS

H. C. Hawkins
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

The performance of frequency shift radioteletype converter equipment is determined by a number of factors. A series of measuring procedures for determining the limitations imposed by any single one or combination of these factors is described. Included in the tests are measurements of signal to random and impulse noise, tolerable input signal drift, interference susceptibility, input level range, magnitude of frequency shift, diversity action, and keying speed limits in terms of teletype printer hits and telegraph distortion. The test procedures described are compared with those performed and reported by other agencies. The need for uniform standards is stressed.

Television I

Transmission Systems

47. 5-Kw VISUAL AND 2.5-Kw AURAL TELEVISION AMPLIFIERS

P. Breen
(Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

This paper describes 5-kw visual amd 2.5-kw aural amplifiers to be driven by commercially available low-power visual and aural television transmitters. The design stresses simplicity in operation and maintenance, since each amplifier consists of only one grounded-grid, push-pull stage using forced air-cooled Eimac 3X2500A3 tubes. Both the visual and aural amplifiers give a power gain of between 5 and 6. Performance data showing compliance with all present and proposed FCC and RMA requirements are given, as well as a complete description of the mechanical and electrical features of both amplifiers.

48. DESIGN CONSIDERATIONS IN TV TRANSMITTERS

L. Pollack, E. Bradburd, and I. Krause
(Federal Telecommunication Laboratories, Inc. Nutley, N. J.)

The following topics of TV transmitter design are discussed and particular solutions presented to the following problems:

(1) The general problem of the TV transmitter, i.e., obtaining wideband amplitude modulation with good linearity, stable characteristics, and dc transmission. (2) The level at which to introduce modulation. The relative advantages and disadvantages of the various methods are discussed in some detail. (3) Selection of the output tube for the aural and visual transmitters. Design of the exciter chains. (4) The modulator is examined with special consideration given to dc stability and economy of design. (5) Stability, regulation, and transient response in relation to power supplies, control, and metering circuits. (6) Sideband filtering from the viewpoint of long-term stability, picture quality, and maintenance of continuous operation. (7) Monitoring methods are presented for properly appraising picture quality and measuring per cent modulation. Methods for measuring aural and visual power outputs are presented. (8) Mechanical considerations, i.e., cooling, ease of maintenance, tube replacement, and tuning.

49. WIDEBAND RF PROBLEMS IN TELEVISION TRANSMITTERS

E. Bradburd and L. Pollack
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

The following topics of TV rf transmitter design are discussed and particular solutions presented: (1) Gain bandwidth products of coupling networks and the application of these principles to power amplifiers; (2) the modulated power amplifier design in relation to such considerations as coupling networks, ease of tuning, long term stability, and drive regulation; (3) phase modulation of the visual carrier and methods for reducing this effect; (4) design of the power output stage with respect to the coupling network, voltage problems, reducing of stray capacitance, effect of long lines on bandwidth, etc.; (5) sideband filter design; and (6) monitoring and general transmitter measurements by the use of reflectometer.

50. THE VIDICON—A NEW PHOTOCONDUCTIVE TELEVISION PICKUP TUBE

P. K. Weimer, S. V. Forgue, and R. R. Goodrich
(RCA Laboratories, Princeton, N. J.)

Photoconductivity offers the possibility of designing a television pickup tube in its simplest, most sensitive, and most compact form. A new television pickup tube called the "vidicon" and employing a photoconductive target is now in an advanced stage of experimental development.

The particular form of vidicon to be described is one inch in diameter and six inches long. It is particularly adapted to industrial applications by virtue of its simplicity and compactness. The target sensitivity is sufficiently high to permit operation at moderate light levels without requiring an electron multiplier. In some cases experimental targets have exceeded one thousand microamperes per lumen.

51. INDUSTRIAL TELEVISION SYSTEM

R. C. Webb and J. M. Morgan
(RCA Laboratories, Princeton, N. J.)

A miniature television camera system has been developed around the new "vidicon" photoconductive pick-up tube for industrial use. A simplified self-contained synchronizing generator in the master control-monitor unit establishes a scanning rate identical with RMA standards for broadcasting, thus making it practical to use existing TV receivers as extension monitors. An improved method of obtaining horizontal deflection in the camera makes possible the use of camera cable lengths up to 500 feet. The camera contains a small video amplifier and a servo motor for remote control of optical focus. Total power consumption is only 350 watts.

Passive Circuits

Filter Circuits and Variable Networks

52. FREQUENCY ANALYSIS OF VARIABLE NETWORKS

Lotfi A. Zadeh
(Columbia University, New York, N. Y.)

The familiar frequency domain techniques commonly used in the analysis of fixed linear networks are extended to linear variable networks. It is shown that a function $H(j\omega; t)$ termed the system function of a variable network, possesses most of the fundamental properties of the system function of a fixed network. It is further shown that $H(j\omega; t)$ satisfies a linear differential equation in t, which is of the same order as the differential equation relating the input and output of the network. Two methods of solution of this equation covering most cases of practical interest are given. On the basis of the obtained results it appears that frequency domain analysis possesses significant advantages over the conventional approach using the impulsive response of the network.

53. DISTORTION BANDPASS CONSIDERATIONS IN ANGULAR MODULATION SYSTEMS

Albert A. Gerlach
(Illinois Institute of Technology, Chicago, Ill.)

The distortion introduced into an angular modulated signal transmitted through a linear network or medium is analyzed using the sideband approach. It is shown that closed form solutions may readily be found for transfer functions which are linear ex-

ponential functions of frequency. A few examples of possible transfer functions are treated illustrating the distortion tendencies for both amplitude and phase characteristics. A rule-of-thumb-formula is derived to determine the maximum undistorted modulation frequency which may be transmitted through a network of a given bandpass characteristic.

54. CONCERNING THE LOWEST POSSIBLE UNLOADED RESONANT CIRCUIT Q'S WHICH CAN BE USED IN MULTIPLE RESONANT CIRCUIT FILTERS

MILTON DISHAL

(Federal Telecommunication Laboratories, Nutley, N. J.)

For the small percentage pass bands (10 per cent to as low as 0.1 per cent in some cases) which are required for the great majority of selective circuit designs in the hf, vhf, and microwave region the usual filter designs produce unnecessarily poor passband response because they assume that the "unloaded Q" of the resonators which are used are very much higher than the fractional mid-frequency ($f_0/\Delta f_{3\,\text{db}}$) being used.

This paper shows how the constants of multiple resonant circuit filters must be varied when the above assumption is not true. In particular it is shown that no loss whatever in sharpness of cutoff need be suffered if the designer can obtain resonators whose unloaded Q is higher than

$$Q_{\min} > \left\{ \frac{1}{\sin\left(\dfrac{90°}{n}\right)} \, \frac{\cosh\left[\dfrac{1}{n}\cosh^{-1}\left(1/\sqrt{(v_p/v_v)^2-1}\right)\right]}{\sinh\left[\dfrac{1}{n}\sinh^{-1}\left(1/\sqrt{(v_p/v_v)^2-1}\right)\right]} \right\} \left(\frac{f_0}{\Delta f_{3\,\text{db}}}\right).$$

A quadruple-tuned 1,500-Mc filter unit is described and shown.

55. TUNABLE MICROWAVE WAVEGUIDE FILTERS

W. SICHAK AND H. A. AUGENBLICK

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

Heretofore, tunable waveguide filters have consisted of cavities connected with quarter-wavelength coupling lines and tuned by means of a lumped reactance in the cavity. Such filters exhibit assymmetrical frequency response and a bandwidth proportional to the cube of frequency when tuned.

A different method of tuning is to maintain constant guide wavelength as the filter is tuned. A filter so tuned will exhibit symmetrical pass-band response and essentially constant bandwidth.

Various methods of varying guide wavelength are investigated, including (1) changing the guide width, (2) inserting a dielectric strip in the broad face of the guide, and (3) inserting a metal strip in the broad face of the guide.

56. FILTERS FOR TELEVISION INTERFERENCE

A. M. SEYBOLD

(Radio Corporation of America, Harrison, N. J.)

Television interference caused by radio amateur transmitters is divided into two classifications: interference produced by transmitters in the form of harmonic radiation, and interference appearing in television receivers as a result of acceptance of strong, amateur-band signals. The amateur's solution of these problems is discussed.

Shielding the transmitter, filtering its supply and control leads, and inserting a low-pass filter in the transmission line minimize harmonic radiation.

High-pass filters placed at television-receiver antenna terminals improve rejection of amateur signals. Simplified equations for low-pass filters and both shunt-derived π-section and series-derived T-section high-pass filters are presented.

Antennas II

57. ON THE RELATION BETWEEN THE GEOMETRY AND THE IMPEDANCE CHARACTERISTICS OF TYPICAL RADIATING SYSTEMS

T. H. CROWLEY AND V. H. RUMSEY

(Ohio State University Research Foundation, Columbus, Ohio)

A systematic experimental investigation of the effect of the geometry of a radiating system on the impedence versus frequency characteristic, based on Schelkunoff's analysis, leads to a procedure for controlling the impendence characteristic by means of the geometry. The particular cases of a radiator in free space and in the presence of a plane reflector are considered in detail. The general principles of frequency compensating apply to wire antennas, slot antennas, and radiators in cavities. The frequency compensating effects obtainable in practice are strikingly borne out by measurements.

58. A METHOD FOR STUDYING THE RESPONSE OF LOOPS TO THE ELECTROMAGNETIC FIELD

BEVERLY C. DUNN, JR.

(Harvard University, Cambridge, Mass.)

Loop response is studied by means of a cavity resonator of specialized design. The loop is placed in a rotary mount fixed in position along a tube equipped with identical pistons. The pistons can be moved axially and rotated; thus by mounting exciter, monitor, and polarizer on the pistons, a resonant mode can be moved axially and rotated with respect to the loop position by corresponding motions of the pistons. A wide range of complexity of loop excitation is therefore available for investigation of loop response to intricate field distributions as a function of the characteristics of the loop system.

59. MEASUREMENT OF THE RADIATION EFFICIENCY OF ELLIPTICALLY AND LINEARLY POLARIZED ANTENNAS

JOHN ROWEN

(Ohio State University Research Foundation, Columbus, Ohio)

The development of a direct reading efficiency meter has made possible a study of the radiation efficiency of a number of linearly and elliptically polarized antennas. The operation and calibration of the efficiency meter will be described briefly, and verification of its accuracy will be made upon the basis of measurements of a number of substantially lossless antennas. The use of this meter on a number of polyrods and several elliptically polarized antennas will be discussed.

60. BROAD-BAND UNDIRECTIONAL ANTENNA 50 TO 170 MC

V. J. COLAGUORI AND R. GUENTHER

(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

This paper deals with the design and test of a dipole array consisting of two driven cylindrical elements relatively large in diameter with suitable reflectors. The required frequency range is 50–170 Mc (3.4 to 1), the nominal input impedance 50 ohms, unbalanced, with deviations within about 3:1 VSWR. The pattern requirements call for a unidirectional characteristic with side lobes, irrespective of their direction, 10 db below the desired direction. Their requirements represent just about the limit realizable with a dipole array. Calculations of the dimensions of the antenna and balancing transformer are carried out. Measurements show the achieved characteristics of the antenna and balancing transformer separately, as well as for the entire system.

61. ANTENNA SYSTEM FOR VERY-HIGH-FREQUENCY RADIO RANGES AND DIRECTION FINDING

F. J. LUNDBURG AND F. X. BUCHER

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

A cylindrical bird-cage like structure which is excited internally by a dipole field and a loop field produces the necessary information for a VOR system. The bird-cage like structure modifies the impedance of the short rotating dipole to provide a suitable match over the frequencies assigned to VOR's. The cage structure minimizes electric current sheet radiation resulting in a high degree of pure horizontal polarization.

Since the system utilizes a rotating dipole for azimuth information, errors associated with systems employing goniometers are eliminated. Space is available for suitably mounting distance measuring equipment giving a complete ρ, θ system.

Systems III

Modulation Systems and Bandwidth Requirements

62. PRODUCT PHASE MODULATION AND DEMODULATION

D. B. Harris and D. O. McCoy
(Collins Radio Co., Cedar Rapids, Iowa)

A type of phase modulation system is described in which modulation of the phase angle of a carrier wave is brought about through an amplitude modulation process. The modulating voltage e_m, is first impressed in parallel on a "Sine Phase Converter" and a "Cosine Phase Converter," which produce respectively, at their outputs, the functions sin ke_m and cos ke_m. The first of these modulating functions is used to modulate, in a balanced modulator, the carrier wave, cos ωt. The second modulating function modulates in a similar balanced modulator the same carrier wave displaced in phase by 90 degrees, or sin ωt. The outputs of the two modulators are added to produce a phase-modulated wave in accordance with the trigonometric identity sin ke_m and cos ωt+cos ke_m sin ωt=sin $(\omega t+ke_m)$.

The "Phase Converters" proposed consist of oscilloscope tubes in which the phosphor has been replaced with an anode for collecting electrons, and a mask is interposed in the electron beam in order to produce at the anode, a voltage proportional to the sine or cosine of the linear beam deflection, which in turn is proportional to the modulating voltage applied to the deflecting plates. It is indicated that phase deviations as high as $\pm 25\pi$ radians may be obtained with this system.

63. SOME NOVEL METHODS FOR THE GENERATION OF PCM

N. R. Castellini, D. L. Jacoby, and B. Keigher
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

Methods for the generation of pulse code modulation are described which all have in common the characteristics that they are self-limiting, self-quantizing and that they can be operated at relatively high speed by trading bandwidth for time. The circuits are relatively simple in that the number of elements in the circuits is proportional to the number of digits and not to the number of levels. In these methods the code is obtained by operating directly on pulse-width modulation by using delay lines as yardsticks of comparison. In addition to being useful for the generation of binary PCM, the methods are readily adaptable to the generation of PCM to higher base (i.e., ternary, quaternary, etc.). An experimental circuit designed to operate at 1-Mc repetition frequency will be described.

64. A HIGH-CAPACITY MATRIX-COMMUTATED RADIO TELEMETERING SYSTEM

J. P. Chisholm, E. F. Buckley, and G. W. Farnell)
(Massachusetts Institute of Technology, Cambridge, Mass.)

Discussion of basic requirements for high capacity pulse-time telemetering system. Development of a simple matrix commutator capable of providing 128 sequential channels with as many as 2,000 information samples per second on each channel. Description of a subminiturized PAM-FM telemetering system employing a matrix commutator and a reactance-switched frequency-modulated transmitter. Discussion of applicability of these techniques to voice communication systems.

65. TECHNIQUES FOR CLOSER CHANNEL SPACING AT VHF AND HIGHER FREQUENCIES

Charles F. Hobbs and Walton B. Bishop
(Air Force Research Laboratories, Cambridge, Mass.)

Physically separated radio transmitters can be frequency-division multiplexed to obtain a great economy of bandwidth utilization at vhf and higher frequencies by use of a common reference signal in lieu of locally generated high-frequency oscillations. In an individual transmitter, a modulated low-frequency sub-carrier is heterodyned with a harmonic of the reference signal, and the sum (or difference) frequency band is selected for transmission. Amplitude-modulated voice signals spaced less than 10 kc apart at greater than 100 Mc have been separated with ease by use of a standard low-frequency communication receiver preceded by a wide-band mixer using a local oscillator signal synchronized with the transmitter reference signal.

66. HIGHLY SELECTIVE MOBILE RECEIVER

R. T. Adams and R. A. Felsenheld
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)
G. W. Sellers
(Capehart-Farnsworth Corporation, Fort Wayne, Ind.)

A mobile receiver for operation between 152 and 174 Mc is described that permits reception where adjacent-channel (60 and 120 kc away) interference is of the order of volts. Its selectivity curve is 30 kc wide at the 6-db points and at 30 kc off resonance the attenuation is 100 db. High overload capabilities of the input circuits withstand interfering signals 100 db above the desired signal without loss of sensitivity. The distortion at high signal levels is low. Novel circuit elements and techniques used in the receiver will be described.

Industrial Instruments

67. USE OF IMAGE CONVERTER TUBE FOR HIGH-SPEED SHUTTER ACTION

A. W. Hogan
(Naval Ordnance Laboratory, Silver Spring, Md.)

The equipment described provides a means of obtaining high-speed photographs while utilizing a continuous light source. The paper will describe apparatus for "one shot" exposures of one microsecond or more, and will discuss motion picture and stroboscopic applications of the scheme.

The heart of the equipment is an image phototube such as type 1P25. Images are impressed on the photocathode and the tube is pulsed electrically for a duration equal to the exposure time desired. The image will then appear on the fluorescent screen, and may either be viewed directly or photographed.

68. ULTRASONIC PULSE INSTRUMENTS FOR AUTOMATIC CONTINUOUS MEASURMENT OF PHYSICAL PROPERTIES OF SOLIDS AND LIQUIDS

Stanley R. Rich
(Raytheon Manufacturing Co., Waltham, Mass.)

Ultrasonic instrumentation using the properties of transmission of ultrasonic pulses is used to measure the elastic properties of solids and the viscosity of liquids. The time of propagation of ultrasonic pulses is solids is used in an automatic recording instrument to yield a continuous record of modulus of elasticity showing the changes that occur as the material is put through programs of variation in stress, elongation, temperature and humidity. The attenuation of ultrasonic pulses in liquids is measured and recorded in an automatic recording instrument to measure and control viscosity of oils and other liquids and the polymerization of plastics.

69. ELECTRONIC DUPLICATOR ATTACHMENTS FOR AUTOMATIC MACHINE TOOL OPERATION

Wilfred Roth
(Raytheon Manufacturing Co., Waltham, Mass.)

Servomechanisms employing electronic control elements designed for attachment to existing machine tools to permit the automatic duplication of given patterns are discussed. A single-motion duplicator, when attached to a lathe, is used to automatically turn figures of revolution and, when attached to a milling machine, is used for die sinking. A two-motion duplicator, in conjunction with a standard miller, is used to reproduce contours of planar figures. By proper utilization of these two basic systems, three-dimensional patterns can be automatically duplicated. Discussion of the design requirements, stability criteria, and actual electromechanical embodiments are presented.

70. MODERN METHODS OF SERVO SYNTHESIS

Rawley McCoy and Donald Herr
(Reeves Instrument Corporation, New York, N. Y.)

Cauer, Gewertz, Fry, Bode, Foster, and Guillemin, among others, have made material contributions in the field of network analysis and synthesis pertaining to physical realizability, synthesis to a prescribed frequency function, equivalence, and duality. Many of these results are directly applicable to the analysis and design of servomechanisms, especially to the design of *RC* networks and amplifiers in the control loops, and to the problem of system stability. Noise, nonlinearity, and departure from ideal of servo components, and the techniques of deter-

mining the required networks, their functional forms, and their interconnections, are remaining problems of servo design. The use of one tool, a flexible differential analyzer of the analogue type, in modern servo design, is described; the proposed use of a second tool, a flexible polynomial-plotter, in modern servo design is also presented. In a typical case economically justifying such techniques, their application to a specific system design problem to fulfill prescribed steady-state and transient performance requirements, is made.

71. AN ELECTRONIC FLOWMETER AND ITS INDUSTRIAL APPLICATIONS

EUGENE MITTELMANN AND V. J. CUSHING
(Mittelmann Co., Chicago, Ill.)

Based on the earlier work of A. Kolin, an electronic flowmeter for industrial applications was developed. If the flowing liquid passes through a magnetic field, a potential proportional to the velocity of the liquid will be developed across two measuring electrodes. It is shown that the velocity potentials are within all practical limits independent of the distribution of the laminar flow and also within wide limits of the conductivity of the liquid. The amplifier and interference compensation circuits are discussed, along with various uses for the new flowmeter in the chemical, food, and electronic industries.

SYMPOSIUM

Television

72. A round-table discussion of current television developments, including such items as color television and uhf allocation. The session will be presided over by Donald G. Fink as moderator.

Antennas III

(This session has been canceled due to conflict with the Television Symposium)

73. RADIATION FROM CIRCULAR CURRENT SHEETS

W. R. LEPAGE, C. S. ROYS, AND S. SEELY
(Syracuse University, Syracuse, N. Y.)

An analysis is carried out for the three-dimensional radiation pattern of a system of coplanar concentric cylindrical current sheets. The cylinders are of small height, and have elements perpendicular to the plane of the circles. Two solutions are available; one is general and the other emphasizes beam formation. A prescribed horizontal pattern can by synthesized by the use of either solution. In one case the pattern is expressed as a Fourier series, and the array reduces to a single circle. In the other case the pattern is expanded in a Bessel-Fourier series, and a system of concentric circles is required. The two methods may be combined to simultaneously synthesize prescribed horizontal and vertical patterns.

74. RADIATION PATTERNS OF CIRCULAR AND CYLINDRICAL ARRAYS

JOHN E. WALSH
(Air Force Research Laboratories, Cambridge, Mass.)

Approximate formulas are derived, in closed form, for the horizontal and vertical patterns of arrays of dipoles arranged on an arc of a circle, or on the surface of a right circular cylindrical segment. Patterns are discussed from the point of view of dependence of side lobe level and beamwidth on the radius of the arc or segment and the central angle subtended.

75. PROPERTIES OF GUIDED WAVES ON INHOMOGENEOUS CYLINDRICAL STRUCTURES

RICHARD ADLER
(Massachusetts Institute of Technology, Cambridge, Mass.)

An analysis is given of some basic properties of exponential modes on passive cylindrical structures in which ϵ, μ, and ρ vary, over the cross section. The bounding surface need not be completely opaque; open structures (like the dielectric rod) and those bounded by a wall with prescribed admittance are considered. Major, but not exclusive, attention is given to nondissipative systems. The modes are compared to those encountered in conventional waveguides with respect to orthogonality conditions, power-flow properties, characteristics of the propagation constant, and frequency-dependence of the field configurations.

76. SCATTERING OF PLANE ELECTROMAGNETIC WAVES BY A PERFECTLY CONDUCTING HEMISPHERE OR HEMISPHERICAL SHELL

EDWARD KENNAUGH
(Ohio State University Columbus, Ohio)

Expressions are derived for the total field produced when a plane polarized electromagnetic wave is incident upon a perfectly conducting hemisphere or hemispherical shell. No assumptions limiting the ratio of diameter to wavelength are necessary. Evaluation of related coefficients is shown to be dependent upon the solution of an infinite set of linear equations, which can be ap-approximated with sufficient accuracy by solving a finite subset. For a ratio of radius to wavelength of 0.376 and incidence along the symmetry axis, diffraction by a hemispherical shell is treated by this method and the total field obtained.

77. DIFFRATION BY A PROLATE SPHEROID

F. V. SCHULTZ
(University of Michigan, Willow Run Airport, Ypsilanti, Mich.

The problem of the diffraction of a plane electromagnetic wave by a perfectly conducting prolate spheroid is solved for the case in which the incident wave strikes the spheroid nose-on. The solution involves setting up for the diffracted wave a series in terms of two sets of solutions of the vector Helmholtz equation. The undetermined coefficients used in this series are evaluated by using the boundary conditions existing on the surface of the spheroid. Numerical results are found for the back-scattering cross-section of a spheroid with a particular eccentricity, and these results are compared with those obtained using physical optics and geometrical optics.

Television II

UHF and Color TV

78. A 1-Kw UHF TELEVISION TRANSMITTER

T. M. GLUYAS
(Radio Corporation of America, Camden, N. J.)

A 1-KW uhf television transmitter operating in the channel from 529–535 Mc is described. Methods of employing eight standard tubes in the output stage to develop the required power, design problems encountered, the measuring techniques employed, and novel features are covered. Performance data is included. Photographs of oscilloscope and kinescope monitors accompany the paper.

79. A SUPERGAIN UHF TELEVISION TRANSMITTING ANTENNA

O. O. FIET
(Radio Corporation of America, Camden, N. J.)

A uhf television transmitting antenna of high power gain will be described. This antenna is designed to operate in the channel from 529 to 535 Mc. Problems associated with transmission lines and the application of waveguide techniques to the elements of the radiating system will be discussed. Performance data, design problems, experimental techniques, and novel features will be presented. Illustrations will accompany the paper.

80. DESIGN OF A HYBRID RING DIPLEXER FOR ULTRA-HIGH-FREQUENCY TELEVISION USE

W. H. SAYER AND J. M. DE BELL, JR.
(Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

The hybrid ring diplexer solves two of the problems which beset uhf transmitter designers: (1) In one television channel it permits the mixing of sound and video frequencies with substantially no interaction, an the same time providing feeding outputs for a turnstile type of antenna. (2) It allows the superposition of one or more pairs of synchronized uhf power sources for attaining a high-carrier power from inherently low-powered units.

This paper will outline the simple design considerations, and discuss the performance and applications of a hybrid ring diplexer at 610 Mc.

81. CONSTRUCTION AND OPERATION OF AN EXPERIMENTAL UHF TELEVISION STATION

RAYMOND F. GUY AND FREDERICK W. SMITH
(National Broadcasting Co., New York, N. Y.)

The engineering considerations involved in the construction and operation of an experimental television broadcast station in the ultra-high-frequency band are presented. The transmitter, KC2XAK, is located in Bridgeport, Conn., and operates in the band from 529 to 535 Mc with a newly developed high-gain antenna.

82. ELECTRO-OPTICAL FILTERS FOR COLOR TELEVISION

Victor A. Babits and Frank Hicks, Jr.
(Rensselaer Polytechnic Institute, Troy, N. Y.)

An electrically controlled color filter which can be used in different color television systems and which is based on the interference phenomenon of polarized light will be described.

In this color filter the three primary color components of the color picture to be televised are produced by using either the Kerr, the Faraday, or the Pockels effect exhibited by certain substances. These substances used in conjunction with phase shifting and polarizing materials form a device which, under the influence of three different electric or magnetic fields, has three transmission characteristics for the three primary colors. A variety of color control circuits and a single field electro-optical system using these devices for the control of the transmitted primary color components will be described. The application of these electrical color filters to a color television system and to color motion pictures will be presented.

Computers I

Digital Computers

83. STATIC MAGNETIC PULSE CONTROL AND INFORMATION STORAGE

An Wang
(Harvard University, Cambridge, Mass.)

Magnetic materials having rectangular hysteresis loops are utilized to control the transfer of electrical pulse energy through the magnetic core by its state of residual magnetism. A high power controlling ratio is obtainable through the incorporation of unidirectional current devices. A static, high-efficiency pulse power distribution system is then available. By the combination of the information storage property of the residual magnetism and its pulse controlling effect, stored information can be transferred from one magnetic core to another by electrical pulse only. Digital information can thus be stored, read-out, and shifted between cores statically, constituting a new memory system.

84. THE DESIGN OF DIODE GATE CIRCUITS

Ralph J. Slutz
(National Bureau of Standards, Washington, D. C.)

A generalized procedure of designing diode circuits for pulse gating has been developed with particular attention to the extremely high reliability required for electronic digital computers. To insure this reliability the procedure guarantees a specified minimum performance under the worst possible combination of tolerance extremes for the diodes, resistors, supply voltages, and circuit capacitances. Circuits have been successfully included in the computer under construction at the National Bureau of Standards in which a single vacuum tube is driven by a gating complex containing over 40 independent logical elements (diodes). Even larger complexes are clearly feasible if the occasion arises.

85. MARGINAL CHECKING AS AN AID TO COMPUTER RELIABILITY

Norman H. Tayler
(Massachusetts Institute of Technology, Cambridge, Mass.)

Deteriorating components, particularly crystals and vacuum tubes, cause reduction of safety margins and are a principal source of error in digital computing and pulse communication.

Marginal checking varies voltages in logical circuit groups, inducing inferior parts to cause failure while a test program or pulse transmission detects and localizes potential failure. In a digital computer this can be automatically accomplished with the computer itself acting as the detector.

In one trial, the application of this type of preventive maintenance for half an hour per day improved reliability 50 to 1. This technique is applicable to circuits using on-off signals.

86. DEVELOPMENT OF THE CALIFORNIA DIGITAL COMPUTER

David R. Brown and Paul L. Morton
(University of California, Berkeley, Calif.)

In the low-cost general-purpose electronic digital computer being built by the University of California at Berkeley, any one of 10,000 ten-decimal-digit numbers can be obtained from the compact magnetic-drum memory in an average access time of eight milli-seconds. The large memory, storing both orders and data, makes possible much computation without automatic access to input; the memory is loaded from punched paper tape read photoelectrically. Binary-coded decimal numbers pass serially through the arithmetic unit on four parallel channels, and single-address operations (including division and square rooting) will be performed approximately 60 per second.

87. A NEW CLASS OF SWITCHING TUBES FOR DIGITAL APPLICATIONS

Joseph Katz
(University of Toronto, Toronto, Ont., Canada)

Tubes designed to reduce the complexity of electronic switching circuitry are described. The tubes contain one electrode for each input and one for each output channel. The electrodes are designed so that particular combinations of input voltages result in current flow to the corresponding combination of output electrodes. Focusing is provided by a limiting resistor which restricts the total current and hence the cross section of the electron beam. Particular examples of such tubes are Binary Adder, Subtractor, and Matrix Tubes.

These tubes feature cylindrical or conical cathodes, three input electrodes disposed symmetrically around the cathode, and output electrodes placed opposite the windows in or gaps between the input electrodes.

Electron Tubes I

Theory and Design

88. MONOFORMER

A. C. Munster
(Philco Corporation, Philadelphia, Pa.)

The Monoformer which will be demonstrated in the course of this paper is a four-terminal nonlinear device which, by selection of the appropriate Monoformer Tube, may be made to have any desired single-valued transfer characteristic. The Monoformer Tube is a special cathode-ray tube having a target whose geometric properties determine the transfer characteristics of the Monoformer itself. This target is separated into two areas, the line of separation representing a curve which is the transfer characteristic of the monoformer. Signals obtained from the target provide output and are also used to cause the electron beam to rest on the curve independently of the deflection of the beam caused by input signals applied to the other pair of deflection plates.

The response time of the Monoformer is several orders of magnitude faster than motor-driven servo potentiometer systems. Its accuracy is to within 3 per cent.

89. A NEW TYPE OF FREQUENCY-CONTROL TUBE

R. W. Slinkman
(Sylvania Electric Products Inc., Emporium, Pa.)

This paper will cover the development and application of the SR1041A variable capacitance tube. This development proceeded from an idea proposed by Blackstone Associates of Chicago. Variations in bias produce variations in plate current sufficient to actuate a bymetallic element controlling capacity. By shunting this capacity across the oscillator circuit and using the output of an AFC circuit to vary the bias, oscillators of extremely good stability may be obtained. Since the device has thermal lag, it is unaffected by modulation frequencies and has application in FM receivers and transmitters.

90. MIT ELECTROSTATIC STORAGE TUBE

S. H. Dodd, H. Klemperer, and P. Youtz
(Massachusetts Institute of Technology, Cambridge, Mass.)

A beam-deflection electrostatic tube was developed to store binary-coded information at two stable potential levels, 100 volts apart, for digital computers or communications systems. A single 2,000-volt electron beam writes or reads one of 400 binary digits on a four-inch target. A 100-volt electron flood replaces leakage and retains stored information indefinitely. The potential boundary stability on the storage surface is assured by a mosiac of conducting beryllium squares. Access time is 25 microseconds. Tubes are in pilot production for a digital computer. Future developments should increase access speed to 6–12 microseconds and storage density to 1,024 binary digits.

91. PERFORMANCE AND ANALYSIS OF A TRANSVERSE CURRENT TRAVELING-WAVE AMPLIFIER AND LIMITER

L. M. Field
(Stanford University, Stanford, Calif.)

A tube is described in which a wide ribbon beam is sent normal to the direction of propagation of waves on a flattened, skewed helical circuit. As a result waves increase exponentially in both beam and wave propagation directions. The small signal analysis will be summarized and results found to check closely with experiment. Gain in db varies as current squared at low currents, but reaches an asymptotic value at high currents. Typical performance of first models is: net gain 30 db at 200 Mc; operation from 150 to 350 Mc; beam voltage 50 volts; beam current 60 milliamperes; and at limiting levels, less than 0.1 db variation in output for 20 db change in driving power.

92. AN EXPERIMENTAL ELECTRON TUBE USING SPACE-CHARGE DEFLECTION OF THE ELECTRON BEAM

J. T. Wallmark
(RCA Laboratories, Princeton, N. J.)

Conventional electron tubes have been built using either grid control or deflection control; the tube to be described uses a new principle combining both of these methods of operation. The grid-controlled space charge produces displacement of the electron beam on an intercepting edge which is converted into additional anode current variation. Experimental tubes utilizing this principle have been built, incorporating a single-stage secondary-emmission multiplier, and giving a transconductance of 25,000 micromhos with only 3 ma output current. This high transconductance with low current and small capacitance should be make this tube quite valuable for broad-band amplifier applications.

Active Circuits I

Amplifiers

93. THE ANALYSIS AND DESIGN OF A BAND-PASS DISTRIBUTED AMPLIFIER

V. C. Rideout and T. P. Tung
(University of Wisconsin, Madison, Wis.)

An exact analysis of the simple distributed amplifier with $2n$ filter sections and $n-1$ tubes has yielded general expressions for gain and phase. These expressions when applied to the band-pass case show that the midband gain is increased by 20 log $(n-1)$ decibels over the single-tube case, and also that some ripple and peaking at the edges of the band are unavoidable by any simple means.

Parallel-tuned transformer filter sections are shown to have some advantages over constant-k filter sections in the band-pass case, and the use of series-parallel transformers as terminating sections gives further advantages. A four-tube 6AK5 amplifier of this type has been built and gives 15 db gain over a 23 Mc band centered at 65 Mc.

94. AN INVESTIGATION OF THE 400-MEGACYCLE AMPLIFIER PERFORMANCE OF THE TYPE SN-973B SUBMINIATURE RF PENTODE

Norman B. Ritchey
(Sylvania Electric Products Inc., Kew Gardens, L. I., N. Y.)

The design of a single-stage rf amplifier is discussed. This unit was built to evaluate the performance of the subminiature pentode at ultra-high frequencies.

The measurement techniques used to evaluate tube impedances are outlined. From these impedances and conventional theory on impedance matching, the stage gain was evaluated. Measurements were made on an amplifier stage and a comparison is given, showing that performance predictions are possible at these frequencies. Measurement techniques used in evaluating tube gain and bandwidth are also given.

95. AN EXTENSION TO STAGGER-TUNED AMPLIFIER DESIGN

Joseph M. Pettit
(Stanford University, Stanford, Calif.)

This paper concerns stagger tuning of amplifier stages having double-tuned coupling networks rather than the customary single-tuned circuits. The former networks have substantially greater gain-bandwidth factors.

One previous approach is Wallman's "stagger damping", which has been treated for the *narrow-band* (or high-Q) case only. By use of a new lowpass-bandpass transformation, arbitrary bandwidths can be accomodated and greater flexibility is afforded in the design of the individual circuits. A specific amplifier is described which has three double-tuned stages *staggered* to provide an over all fractional bandwidth of 0.67. The bandwidth is almost 30 per cent greater than would be provided by three identical stages with the same over-all gain. The difference would be even more pronounced with a greater number of stages.

96. ULTRA-HIGH-GAIN DIRECT-COUPLED AMPLIFIER CIRCUITS

W. K. Volkers
(Consulting Engineer, Schenectady, N. Y.)

By lowering the screen voltage of pentodes below 10 per cent of their plate supply voltage and increasing the resistance of their plate load 10 or more times beyond conventional values the amplification factor of tubes, "starved" in this manner, is greatly increased in spite of a decrease of their mutual transconductance. Stage gains as high as 2,500 were measured in this condition, using regular 6SJ7 tubes. By incorporating the principle of starvation in a new direct-coupled amplifier two basic aims can be achieved: (a) a drastic increase of over-all gain permitting omission of amplification stages and reduction of manufacturing costs (example: 3-tube radio receiver having only 4 resistors and 4 condensers), or, (b) "trading" of surplus gain for minimum distortion and maximum stability (examples: TV oscilloscope amplifier and sensitive VTVM for dc).

97. ANALYSIS AND DESIGN OF SELF-SATURABLE MAGNETIC AMPLIFIERS

S. Cohen
(Sperry Gyroscope Co., Great Neck, L. I., N. Y.)

A self-saturable magnetic amplifier circuit consisting of a reactor winding in series with a dry disc rectifier, a resistive load and an ac voltage source is described and analyzed.

Optimum values of load resistance and power can be obtained as function of the quality of core material (*B-H* curve as approximated by two straight lines), type of core, and characteristics of the rectifier elements. A fair degree of agreement is obtained with experimental results.

Amplifier stages made up with a combination of these basic circuit elements can be used as multistage amplifiers in servomechanisms and other applications.

SYMPOSIUM

Basic Circuit Elements

98. PERFORMANCE—MEASUREMENT OF CAPACITORS

H. T. Wilhelm
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

The performance of capacitors as a function of frequency will be presented with particular emphasis on effects of dielectric loss and inductance. Established measurement concepts and techniques will be discussed and limitations of the existing methods of measurement will be pointed out; in particular, the residual errors and connection errors. Special measurement techniques such as insertion and substitution methods will be described.

99. BEHAVIOUR OF RESISTORS AT HIGH FREQUENCIES

G. R. Arthur, H. L. Krauss, and P. F. Ordung
(Yale University, New Haven, Conn.)
S. E. Church
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

Results are presented on recent investigations for the behavior of carbon resistors as a function of frequency up to 50 Mc. Using a distributed-constant equivalent circuit, the prediction of impedance values from low-frequency measurements is possible. For wide-band applications, a distributed constant resistor which utilizes capacitance to ground as a parameter is under development, and preliminary test results will be given.

100. INDUCTORS, THEIR CALCULATION AND LOSSES

Robert F. Field
(General Radio Co., Cambridge, Mass.)

Methods of calculating the inductance of both air- and iron-cored inductors will be presented, including some refinements with

particular emphasis on distributed capacitance. The treatment of losses will be based on the dissipation factor, and will include single layer and multiple layer solenoids and toroids. The importance of the dissipation factor of the distributed capacitance will be stressed.

101. TRANSFORMER PERFORMANCE AND MEASUREMENTS

Reuben Lee
(Westinghouse Electric Corp., Baltimore, Md.)

Performance characteristics of transformers as a function of frequency will be discussed with particular emphasis on the influence of distributed capacitance in the higher frequency ranges. The methods of measuring the response characteristics, the input impedance, and insertion loss will be presented with emphasis on interpretation of the results. Shielding problems, magnetic pickup, and harmonic distortion will be surveyed.

Television III

Receivers

102. USE OF MINIATURE PENTODE RCA-6CB6 IN TELEVISION INTERMEDIATE-FREQUENCY AMPLIFIERS

W. E. Babcock
(Radio Corporation of America, Harrison, N. J.)

New miniature pentode, RCA-6CB6, in video intermediate-frequency amplifiers operating at 41.25 to 45.75 Mc per second offers advantages over comparable types because of lower grid-plate capacitance, higher transconductance, and separate cathode and suppressor connections.

Determination of damping-resistor values for stagger-tuned if amplifiers takes into account input conductance components due to transit-time effects, tube lead inductance, and grid-plate feedback capacitance. Curves show changes in input conductance and capacitance with transconductance. These changes are minimized by use of proper value of unbypassed cathode resistor. Four methods are suggested for improving rejection characteristics.

103. NOISE SUICIDE CIRCUIT

Harold E. Beste and George D. Hulst
(Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

The operation of television receivers in weak signal areas depends upon the ability of the various circuits of the receiver to operate under conditions in which undesired noise has the same order of peak amplitude as the desired signals.

A novel circuit is described which allows the reception of television signals without loss of scanning synchronization under very severe noise conditions. The circuit gives positive discrimination against all common types of undesired noise, including not only short-duration impulses, but also continuous interfering signals, and thermal noise.

104. QUALITY RATING OF TELEVISION IMAGES

P. Mertz, A. D. Fowler, and H. N. Christopher
(Bell Telephone Laboratories, Inc., New York, N. Y.)

In the transmission of television signals, deviations from an ideal medium assume importance. Two techniques were explored for rating the impairments subjectively. Both employ viewing of the received picture by a number of observers. The first is an extension of Baldwin's. The observer votes a choice between two slightly different pictures. The impairment is measured in liminal decrements computed from the vote. In the second the observer characterizes the picture by one of a series of pre-worded comments. The techniques have been used to evaluate the importance of echoes and noise, and make comparisons with sharpness and other quality parameters.

105. TELEVISION IMAGE REPRODUCTION BY USE OF VELOCITY-MODULATION PRINCIPLES

M. A. Honnell and M. D. Prince
(Georgia Institute of Technology, Atlanta, Ga.)

This paper describes a composite television system in which a picture televised by a conventional video camera is reproduced by velocity modulating the horizontal motion of the writing spot on the screen of the kinescope. Photographs are presented which indicate the nature of the image reproduction achieved by velocity television reproduction. This system may offer some advantages in the transmission of printed material and line drawings. An analysis is presented indicating the response of the system to a general image.

106. DESIGN OF PRINTED-CIRCUIT TELEVISION TUNER

D. Mackey and E. J. Sass
(Radio Corporation of America Camden, N. J.)

A high-gain twelve-channel turret-type television tuner has been developed in which the coil system is produced by printed-circuit techniques. Printed coil segments for each channel include a tuned antenna coil, a double-tuned "M"-derived band-pass rf transformer, and a tuned oscillator coil for each of the high-frequency channels.

The factors leading to the choice of the type of printed-circuit technique employed and the comparative advantages of the several methods of producing printed circuits will be discussed.

Other matters of interest in the design of the tuner, including the general mechanical design, the contact design, and the circuitry will be described in some detail.

Active Circuits II

General

107. FREQUENCY-MODULATION INTERFERENCE

L. B. Arguimbau
(Massachusetts Institute of Technology, Cambridge, Mass.)

When two frequency-modulated signals with standard broadcast deviation and preemphasis interfere, the signal-to-interference ratio on the output of an "ideal receiver" is improved by some 30 db with respect to the input ratio. It is shown that with proper receiver design this relationship holds even though the signal and interference differ by only a fraction of a decibel. Methods of computing the root-mean-square interference are discussed. Receiver operation under common-channel and multipath conditions are demonstrated.

108. THE THEORY OF AMPLITUDE-MODULATION REJECTION IN THE RATIO DETECTOR

B. D. Loughlin
(Hazeltine Electronics Corp. Little Neck, L. I., N. Y.)

A mathematical analysis is presented of the AM rejecting properties of the ratio detector. The operation with 100 per cent efficient diodes is first treated and it is shown that in this case compensating resistors which reduce the effective efficiency of the diodes must be used to obtain optimum AM rejection. The operation with practical diodes is then treated and design charts for optimum AM rejection are presented. Apparent limiting action within the ratio detector circuit is described and it is shown that the degree of apparent limiting is incidental and unrelated to the AM rejection properties of the ratio detector.

109. AN IMPROVED METHOD OF FREQUENCY CONVERSION

Vernon H. Aske and J. Grund
(Sylvania Electric Products Inc., Kew Gardens, L. I., N. Y.)

The theory and experimental results are given for a frequency-conversion system which results in a $g_c = 2g_m/\pi$. This compares with the conventional mixer whose g_c approaches g_m/π. A pentode tube is used in a relatively simple circuit to derive these results. This circuit retains the advantages of the conventional pentode mixer while greatly increasing the effective conversion transconductance and overcoming the noise disadvantage of the usual pentode mixer. Comparisons are made between the new circuit and other mixing circuits.

110. COMMON FREQUENCY CARRIER SHIFT RADIO-TELETYPE CONVERTER

R. R. Turner
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

A converter design utilizing a new technique to reduce channel spacing is described. This equipment employs the narrow bandpass filters usually found in the audio converters while retaining the advantages of the IF type of converter. Due to a novel arrangement of circuits, the design would theoretically be less subject to adjacent channel interference, and also permit positive automatic frequency control. Data are presented showing the performance of an experimental model in the presence of random noise, impulse noise, and adjacent channel interference, compared with the standard military converter.

111. A SIMPLE CRYSTAL DISCRIMINATOR FOR FM OSCILLATOR STABILIZATION

J. Ruston
(Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

A crystal discriminator employing only one quartz crystal and very simple in adjustment is described in detail. The input frequency is measured by comparing the crystal impedance with a reference impedance which remains substantially constant of the range of measurement. The accuracy of center frequency measurement in the presence of frequency modulation, and variation of component values can be calculated in terms of the circuit parameters and shown to be adequate for the requirements of broadcast transmitters.

The discriminator was developed for center frequency stabilization of the FM exciter unit in a television sound transmitter. The complete exciter unit is described briefly.

Electron Tubes II

Theory and Design

112. A NEW "SOFT STRUCTURE" FOR RUGGED RECEIVING TUBES TO IMPROVE RESISTANCE TO SHOCK AND ELECTRON EMISSION

George W. Baker
(Kip Electronics Corp., New York, N. Y.)

A rugged receiving tube will withstand a mechanical shock that does not cause any part of it to deflect past its elastic limit. Instead of making the structure of a larger receiving tube very stiff to reduce the deflection during shock, it is designed to deflect considerably and a greater shock can be applied without reaching the elastic limit.

This same design reduces the distortion of the base metal of the cathode during shock, and this has been found to greatly improve the constancy of electron emission. A sample GT type receiving tube will be shown.

113. HYDROSTATIC PRESSURE IN AN ELECTRON GAS: ITS APPLICATION TO ELECTRON BEAM-ELECTROMAGNETIC WAVE INTERACTION

P. Parzen and L. Goldstein
(Federal Telecommunications Laboratories, Inc., Nutley, N. J.)

The concept of hydrostatic pressure in an electron gas has been utilized by Hahn and Bailey to discuss space-charge behavior and electromagnetic wave propagation in electron beams. The feasibility of the application of this concept usually depends upon the degree of validity of the tacit assumption that the process is isothermal. It is shown that the process of electromagnetic wave propagation in an electron gas is approximately isothermal if the ratio of random energy to directed energy is small. Calculations of the amplification of electromagnetic waves in electron beams show that the amplification is less than that due to a monochromatic beam.

114. HIGH-VOLTAGE REGULATION BY MEANS OF CORONA DISCHARGE BETWEEN COAXIAL CYLINDERS

S. W. Lichtman
(Naval Research Laboratory, Washington, D. C.)

The corona discharge between coaxial cylinders affords a practical means for stabilizing high voltages in a manner analogous to the stabilization of low voltages by the familiar glow-tube regulator. The corona regulator is particularly suitable for stabilizing voltages above several hundred volts at currents below one milliampere. It is accordingly well adapted for controlling the beam focusing and accelerating potentials of cathode-ray devices such as oscilloscope, iconoscope, and kinescope tubes, of electron diffraction cameras, and for stabilizing Geiger tube voltage sources. This paper describes some of the important theoretical aspects of the corona discharge region as related to voltage stabilization. Constructional features and performance characteristics of typical high-voltage regulating tubes are described. And circuit design relationships are presented for adapting particular corona regulator tube characteristics to specific performance requirements.

115. THYRATRON GRID EMISSION AND THE TRIGGER-GRID THYRATRON

L. Malter and M. R. Boyd
(RCA Laboratories, Princeton, N. J.)

Electron emission from thyratron grids particularly when increased by gas amplification, limits the performance characteristics. Studies indicate that the thyratron grid serves a double function: (1) as a firing electrode, and (2) as a recovery control device. In many types these functions are combined in a single electrode. By separating the functions, the deleterious effects of grid emission can be reduced. The firing function is assigned to a small area electrode known as the trigger grid, and the recovery function is assigned to another electrode referred to as the blocking grid.

Modification of conventional structures so as to make use of the separation of functions enables the power handling capabilities and sensitivity to be increased several fold. Other accompanying benefits are: (1) decreased recovery time with large values of grid resistor; (2) decreased plate-grid capacitance; and (3) less critical heater voltage.

116. HIGH-INTENSITY PULSE-DISTRIBUTION TUBE

P. M. G. Toulon
(Consulting Engineer, Neuilly-on-the-Seine, France)

Electronic processes for gating a multiplicity of channels at low level are well known, but there is need for a device capable of providing distribution of current pulses of relatively high intensity.

A gas tube having as many as 64 anodes has been used for distribution of impulses of two amperes peak. With another special tube, peak currents of over 100 amperes have been commutated.

An application of this device in conjunction with a master oscillator synchronized with the incoming signal is described.

Computers II

Information Analysis and Computing

117. A DISCUSSION REVEALING SOME LATE DEVELOPMENTS IN ELECTRONIC ANALOG COMPUTER TECHNIQUES

H. I. Zagor
(Reeves Instrument Corp., New York, N. Y.)

The REAC and associated computing components as a flexible and practical tool for solving ordinary linear and nonlinear differential equations will be described. A discussion of the capabilities of the various components such as amplifiers, integrators, servos, resolvers, input-output tables, limiters and relay amplifiers will be given. Illustrative problems in such diverse fields as flutter, electron flow, automatic pilot design, Fourier analysis, engine control, integral and boundary value equations will be presented and various techniques involved in obtaining these solutions will be shown.

118. AN ELECTRONIC STORAGE SYSTEM

E. W. Bivans and J. V. Harrington
(Air Force Research Laboratories, Cambridge, Mass.)

A digital storage system using the RCA Radechon, a barrier grid storage tube, is described. The information to be stored consists of code groups of 9 pulse positions in a 10-μsec interval. A novel method of Radechon operation is employed in that the secondary collector system is not used, instead, the reading signals are measured at the back plate, the same electrode on which the write signals are impressed. Deflection voltages are generated by a weighted addition of the plate voltages of a binary counter. Read and write operations are asynchronous with a 12-μsec minimum time between operations.

119. EXPERIMENTAL DETERMINATION OF SYSTEM FUNCTIONS BY THE METHOD OF CORRELATION

J. B. Wiesner and Y. W. Lee
(Massachusetts Institute of Technology, Cambridge, Mass.)

An application of the theory of correlation functions is made in the determination of the transfer functions of linear and nonlinear systems. Random noise with a uniform power density spectrum over a wide range is used as the source of power. The auto-correlation function of the input and output are obtained by means of an electronic correlator. System functions are then determined from the correlation functions. Illustrative cases are presented.

120. MEASUREMENT AND ANALYSIS OF NOISE IN A FIRE-CONTROL RADAR

R. H. Eisengrein
(Sunstrand Machine Tool Co., Rockford, Ill.)

The determination of an optimum design for an automatic-tracking radar-controlled fire-control system is hindered by a serious problem of analyzing the character of the radar noise, i.e., the unwanted portion of the radar signal return from an airborne target. This paper discusses an "instantaneous subtraction" method of optically measuring radar noise, and its value is assessed on the basis of flight test data. The procedure for deriving a noise power spectrum from the calculated autocorrelation function of a noise record is described. Typical autocorrelation and power spectrum curves are presented.

121. A DIGITAL ELECTRONIC CORRELATOR

H. E. Singleton

(Massachusetts Institute of Technology, Cambridge, Mass.)

Earlier experimental work has demonstrated the practical utility of electronic computation of correlation functions, and has led to the design of an improved and more flexible electronic correlator. The new machine accepts inputs covering a wide frequency range (dc to 10 megacycles) and evaluates correlation functions for arguments from 0 to 0.1 second. In order to obtain a high degree of accuracy and stability, the signals are sampled and converted to binary numbers, and the storage and computation are carried out digitally. The theory, design, and application of the correlator are discussed, and a number of results are presented.

Transmission and Antennas

122. SURFACE-WAVE-TRANSMISSION LINES

George Goubau

(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

In 1899, A. Sommerfeld published a paper about wave propagation along a cylindrical wire of finite conductivity. The type of surface wave he investigated has found little consideration. As a matter of fact, the field extends far from the conductor and it may have appeared questionable whether such a wave could be excited and used for transmission lines. However, it can be shown that coating of the conductor by a thin layer of dielectric or any other modification of the conductor surface which is proper to reduce the phase velocity results in a shrinking of the cross section of the field. Such modified surface waves can be easily excited. Transmission lines using these waves have been built and examined for their applicability for microwaves. The measured transmission loss is a fraction of that in coaxial lines, and approaches the loss in waveguides.

123. FREQUENCY-MODULATION DISTORTION IN LINEAR SYSTEMS HAVING SMALL SINUSOIDAL IRREGULARITIES IN TRANSFER CHARACTERISTICS, WITH APPLICATION TO LOSSLESS WAVEGUIDES

F. Assadourian

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

A first-order analysis yields formulas for harmonic and total distortions for pure frequency modulation at a single frequency in linear transmission systems having transfer amplitude and phase characteristics departing from flatness and linearity, respectively, by small sinusoidal variations.

Total distortion varies linearly with the amplitude of either sinusoidal variation if the other is made zero. The distortion formulas are applied to lossless waveguides terminated by pure resistances. The dependence of total distortion on such parameters as waveguide length and modulation frequency is illustrated numerically and graphically. The distortion formulas may also be applied to amplifiers and filters with suitable transfer characteristics.

124. THE REPRESENTATION, MEASUREMENT, AND CALCULATION OF EQUIVALENT CIRCUITS FOR SLOTS IN RECTANGULAR WAVEGUIDE

J. Blass, L. Felsen, H. Kurss, N. Marcuvitz, and A. A. Oliner

(Microwave Research Institute, Polytechnic Institute of Brooklyn, Brooklyn, N. Y.)

The equivalent circuits representative of microwave structures are usually critically dependent upon the choice of reference planes. "Invariant" representations will be presented in which the values of the circuit parameters depend solely on electrical measurements (independent of any absolute distance measurements), while the less accurate mechanical measurements affect only the location of the reference planes. A general description and some details of the precision measurement methods employed will also be given. The measurement methods and the equivalent circuit representations will be discussed in connection with a systematic investigation of the impedance properties of rectangular slots in rectangular waveguide. The theoretical program undertaken in connection with this investigation utilizes variational formulations for aperture-type discontinuities. The appropriate variational expressions will be given and briefly discussed. Experimental and theoretical results for a wide range of slot dimensions will be given for transverse slots coupling two similar guides, slot coupled E-plane Tee's, and slots radiating into space from the end and from the broad face of the guide. Effects of thickness will also be considered.

125. DIELECTRIC TUBE ANTENNAS

R. E. Beam and D. G. Harman

(Northwestern University, Evanston, Ill.)

Experimentally determined radiation patterns of dielectric tube antennas of uniform, tapered and flared cross sections are given for various tube lengths and ratios of inside-to-outside diameters including solid rods. Curves representing the variation of the most important features of the radiation patterns obtained from over 200 experimentally determined patterns are plotted for tubes excited in the hybrid HE_{11} mode. An approximate end-fire array theory of the radiation patterns is given. Results indicate that dielectric tube antennas have desirable characteristics for use in highly directive antenna arrays. Dielectric tubes having ratios of inside-to-outside diameters of 0.7 or greater have side lobes which are 20 decibels below the major lobe.

126. MEASUREMENT OF CURRENT AND CHARGE DISTRIBUTIONS ON ANTENNAS AND OPEN-WIRE LINES

D. J. Angelakos

(Harvard University, Cambridge, Mass.)

An experimental investigation has been made of the distributions of current and charge amplitude and phase on cylindrical antennas and on the driving lines. The presence of a stub support modifies the distributions on the antenna only near the junction of the antenna, line, and stub; however, for antenna loads near antiresonance, the distributions on the line as well as the impedances of the structure are altered considerably.

A corrective network is defined for the terminal region of the junction. Impedances of theoretical models of antennas, the network, and conventional transmission-line equations may then be used to predict the apparent impedances of the structure.

Audio

Transducer Design

127. SOUND SYSTEM DESIGN FOR REVERBERANT AUDITORIUMS

L. L. Beranek, W. H. Radford, and J. B. Wiesner

(Massachusetts Institute of Technology, Cambridge, Mass.)

In connection with The Mid-Century Convocation at MIT, sound systems were installed in three auditoriums whose volumes range from 237,000 to 5,500,000 cubic feet, and in which there is no acoustical treatment. Calculations based on reverberation data taken in these auditoriums and on articulation index theory revealed the need for over-all frequency characteristics which were substantially flat from 500 to 3,000 cps, and which were tailored in such a way, both above and below those frequency limits, to reduce the effects of reverberation and to preserve naturalness of speech. Audience reactions indicated near-perfect intelligibility and little deterioration in voice naturalness.

128. HIGH-EFFICIENCY LOUDSPEAKERS FOR PERSONAL RADIO RECEIVERS

H. F. Olson, J. C. Bleazey, J. Preston, and R. A. Hackley

(RCA Laboratories, Princeton, N. J.)

The term "personal radio receiver" is used to designate a radio receiver with self-contained power supply, and one small enough to be easily carried by hand or in the pocket. Performance and compactness of such receivers is limited by the efficiency of conversion from electrical power to sound power by the loudspeaker. Since electrical power output is limited in such receivers,

loudspeaker efficiency is an important factor. Theoretical and experimental investigations have been made of direct radiator, combination direct radiator and phase inverter, horn, and combination horn and phase inverter type loudspeakers. An efficiency of 25 per cent has been obtained with the combination horn and phase inverter. This loudspeaker system has been incorporated in a complete 4-tube radio receiver having a content of 25 cubic inches.

129. A REVIEW OF DIRECT RADIATOR LOUDSPEAKERS

F. H. Slaymaker
(Stromberg-Carlson Co., Rochester, N. Y.)

Loudspeakers are multiple resonant structures having irregular frequency response and transient characteristics. These characteristics can be controlled by suitable choice of diaphragm material, diaphragm shape, flux density, and amplifier impedance. Other loudspeaker defects include the generation of subharmonics and rattles. Rattles, however, are often attributable to minute amounts of clipping, and short-lived oscillations in the amplifier too small to be measured with conventional distortion meters. An air of mysticism is often associated with loudspeakers and, although this paper is not intended to be definitive, it is hoped that it will help put the choice of loudspeakers on a realistic basis.

130. LOUDSPEAKER HOUSINGS

Willard F. Meeker
(Stromberg-Carlson Co., Rochester, N. Y.)

The performance of even the best loudspeaker may be greatly degraded by the use of an inadequate housing. The usual open-back cabinet of present-day radio and television receivers invariably introduces undesirable resonances. Closed-back and phase-inverter housings are often used in more elaborate reproducing systems. The recent application of room acoustics theory to loudspeaker housings necessitates a revision of the conventional theory for both the closed-back and the phase-inverter housings. The problem of evaluating the performance of loudspeakers in these housings has not been completely solved. Listening tests may be misleading unless the loudspeaker position in the listening room is considered.

131. A MINIATURE CONDENSER-TYPE MICROPHONE

John K. Hilliard
(Altec Lansing Corp., Hollywood, Calif.)

This paper describes a microphone system, the acoustic transducer consisting of a condenser unit having an over-all diameter of 0.6 inch. Factors governing the size of a microphone and how it is related to its frequency response are discussed. The general design considerations are explained. The microphone is omnidirectional over a 360° spherical pattern, has high sensitivity and is extremely rugged to shock. All of the special accessories are discussed in relation to their use in broadcasting, television, and recording application. A short demonstration will follow the paper, showing the various characteristics of the microphone.

Electronics in Medicine

132. EFFECTS OF INTENSE MICROWAVE RADIATION ON LIVING ORGANISMS

John W. Clark
(Collins Radio Co., Cedar Rapids, Iowa)

A study has been made of the effects of intense microwave radiation on living organisms, to ascertain whether or not any damage to personnel exposed to such radiation may be anticipated. It was found that certain parts of the body which are not adequately cooled by the blood stream are indeed vulnerable to microwave radiation of the order of one watt per square centimeter. Damage to the eye has been demonstrated in laboratory animals; this damage takes the form of cataract of the lens. Accordingly standards should be established for the protection of personnel who work with high-power microwave equipment.

A mathematical theory has been developed to account for the temperature distribution in tissues which are exposed to microwave radiation and which are cooled by conduction only. This theory checks well with experiment; it predicts that the greatest temperature rise at a given power density will occur at about 10 cm wavelength. With the aid of this theory it is possible to predict the shape of the temperature distribution at any frequency. This is useful in evaluating the benefits or dangers which may be derived from the use of electromagnetic radiations for therapeutic purposes.

133. A DIFFERENTIAL VECTORCARDIOGRAPH

Stanley A. Briller
(Bellevue Hospital, N. Y.)
Nathan Marchand
(New York University College of Medicine, New York, N. Y.)

The electromotive forces generated by cardiac muscle during its activity vary in size and direction. They therefore lend themselves to vectoral representation, and in the case of the human heart all the forces involved may be recorded as a vector-time diagram (vectorcardiogram).

Construction of electronic equipment for making vectorcardiograms has presented several instrumentation problems. One of these has been the problem of avoiding the superimposition of the vectordiagrams of the various phases of the cardiac cycle known as the P form, the QRS form, and the T form. To overcome this and other difficulties, an apparatus was designed and constructed consisting of a dual-beam cathode-ray tube for simultaneous recording of the frontal and sagittal thoracic projections of the spatial cardiac vectors. The design incorporates circuits which automatically achieve separation and photographic registration of the several vector loops of successive heart beats. Other components of the apparatus include: a network to convert heart voltages into an orthogonal set; a three-channel, low-frequency amplifier; a synchronized timing unit; a time marker; an electronically operated camera; and a direct writing interval delineator.

134. ELECTRONIC MAPPING OF THE ELECTRICAL ACTIVITY OF THE HEART AND BRAIN

Stanford Goldman
(Syracuse University, Syracuse, N. Y.)

This paper describes means whereby a picture of the electrical potential distribution on the surface of any desired portion of the human body is shown on the screen of a cathode-ray tube. A number of pickup electrodes are located in an ordered array on the surface of interest. With the aid of an electronic counter type of commutator, the voltages picked up on the electrodes are impressed upon a scanning signal. A radar PPI type of display is then obtained from this signal, wherein the cathode-ray tube screen represents the area of interest, and the light intensity at any point on the screen is proportional to the instantaneous voltage of the corresponding point on the body. In this way a cathode-ray tube screen has been used to show moving pictures of the heart or the brain in action.

It is found that the electrical activity on the surface of the chest gives an informative picture of the conduction of electrical impulses within the heart itself. Results already obtained indicate that the pictures will be useful for diagnostic purposes and will be valuable in studying the physiology of the heart. Electronic mapping pictures of the brain have shown that the alphawaves of the brain are traveling waves and the paths and speed of travel can clearly be seen in the pictures. Insofar as time permits, motion pictures will be shown at the meeting of the heart and brain in action.

Propagation I

Propagation at Ionospheric Frequencies

135. CALCULATION OF EFFECTIVE PHASE, GROUP, AND PULSE VELOCITIES OF WAVE PROPAGATION

A. Fischler, G. H. Sloan, and D. Goldenberg
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

Methods are deduced for calculating the fractional deviations from c (the velocity of light in vacuo) of the effective velocities of unmodulated, simply modulated, and pulse-modulated waves as functions of earth conductivity and dielectric permittivity, frequency, antenna heights, and polarization. The method represents an approach to first-order accuracy of radio ranging over short distances by means of interval timing (e.g., radar, interferometry, etc.). A technique of accurately measuring earth constants is implicitly indicated, even with existing apparatus.

136. AN ATMOSPHERIC WAVE-FORM RECEIVER

William J. Kessler and Sydney E. Smith
(University of Florida, Gainesville, Fla.)

A triggered sweep oscillograph designed expressly for the observation of atmospheric wave forms is described.

A remote omnidirectional antenna is used to feed the signal channel amplifier which must display essentially uniform amplification and delay characteristics throughout the frequency range corresponding to the significant energy components of most atmospheric signals. An 18-microsecond delay line is used in the signal channel to permit delineation of the leading edge of the atmospheric disturbance triggering the sweep generator. The display of time-interval markers is unique in that they appear on a second elevated sweep to eliminate confusion with the characteristic variations of the atmospheric wave forms. This is accomplished with a display tube containing a single electron gun by delaying the marker sweep approximately 1,500 microseconds and simultaneously blocking the signal channel.

An external brightening pulse is provided to brilliance modulate the cathode-ray tube of an associated atmospherics direction finder to permit identification of the azimuth of the recorded wave forms, and to control a solenoid-operated still camera to advance the normally stationary film one frame after each exposure. Typical examples of atmospheric wave forms showing unmistakable ionospheric returns corresponding to a virtual height of approximately 90 kilometers are included.

137. RADIO WAVE PROPAGATION IN A CURVED IONOSPHERE

John M. Kelso
(Pennsylvania State College, State College, Pa.)

Using a double parabola approximation to the Chapman distribution of electron density as a function of height, and the assumption of a curved ionosphere, curved earth geometry, analytic expressions are obtained for the true height of reflection, ray path, reflection coefficient, ground range, and group path. Graphical results are given for the maximum useable frequency factor. Where possible, the above results are compared with results obtained by assuming a plane ionosphere.

All of these calculations are made under the usual restrictions of neglecting the earth's magnetic field, and most of the effects of collisions of electrons with heavy particles.

Electron Tubes III

Power Tubes

138. DEVELOPMENT OF 10-CM HIGH-POWER PULSED KLYSTRON

M. Chodorow, E. L. Ginzton, I. Neilsen, and S. Sonkin
(Stanford University, Stanford, Calif.)

This paper presents the results of a two-year program just completed at Stanford University. The basic problems, both theoretical and practical, which had to be solved during the course of this program will be discussed.

Some of the specific topics to be treated are: operation of klystrons at 400-kv level, design of suitable modulators, pulse transformers, and other components; the design and construction of successful tubes; and the operating characteristics of some of the tubes.

139. SPACE-CHARGE EFFECTS IN REFLEX KLYSTRONS

V. Westberg and M. Chodorow
(Stanford University, Stanford, Calif.)

In simple reflex klystron theory it is usually assumed that the potential in the reflector region varies linearly with distance. For many practical tubes this is not true, due to space charge effects and reflector curvature. The effect of this departure from the linearity in potential can be represented by a numerical factor F which modifies the true transit angle to give an effective bunching angle. The power output, loading, electrical tuning, will all be altered in a corresponding way. By various approximate methods it has been possible to calculate the F factor as influenced by space charge for a wide range of operating parameters. Measurements on some typical tubes agree quite well with the theoretical value. Using these calculated values of the F factor, it is now possible to predict more accurately the operating characteristics of a reflex klystron for any given mode.

140. RECENT DEVELOPMENT IN HIGH-POWER KLYSTRON AMPLIFIERS

C. Veronda and V. Learned
(Sperry Gyroscope Co., Great Neck, L. I., N. Y.)

Early klystons were limited primarily to gridded interaction gaps with ion-focused electron beams. They were limited in power and were not particularly suitable for pulse operation.

Recent work has developed magnetically focused electron beams which are suitable for both pulse and continuous-wave operation at much higher power levels. High-power klystron amplifiers have been developed which have excellent power-gain characteristics, thus making it practical to drive the tubes from stable frequency sources.

The characteristic properties of several types of these new klystron amplifiers will be discussed and typical performance characteristics presented. A comparison of practical performance characteristics, in relation to idealized theoretical characteristics, will be made to show how specific klystrons are useful for particular system applications.

141. A NEW SUPER-POWER BEAM TRIODE

W. N. Parker, W. E. Harbaugh, M. V. Hoover, and L. P. Garner
(Radio Corporation of America, Lancaster, Pa.)

The design and operation of a super-power beam triode tested at 1,000 kilowatts average power input is discussed. A unique system of electron beam forming permits a power gain as high as 1,000 under the conditions required for 80 per cent efficiency. An improved method of anode construction and water cooling has permitted the testing of the tube at an anode dissipation of 500 kilowatts. New mechanical techniques of individual electrode suspension permit accurate alignment of precisely constructed electrodes. Processing and testing procedures are discussed in conjunction with the facilities required for the development and operation of super-power tubes.

142. EXTERNAL CATHODE INVERTED MAGNETRON

Joseph F. Hull
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

In a radically new approach to the generation of high peak-pulsed power an "inverted" magnetron has been successfully operated. This tube consists of a number of "inside out" interdigital magnetron anodes axially stacked together surrounded by a cylindrical cathode whose inner surface is the emitter. Over-all efficiencies greater than 50 per cent have been achieved with peak power outputs in excess of 2-kw, 6 per cent duty cycle pulse operation. Fifty kw at 0.1 per cent duty cycle have been obtained from a tube designed for cw operation.

Navigation Aids

143. ANALYSIS OF COURSE ERRORS IN THE VHF OMNIDIRECTIONAL RADIO RANGE

J. Wesley Leas
(Air Navigation Development Board, Civil Aeronautics Administration, Washington, D. C.)

The Air Navigation Development Board has been evaluating the accuracy of the VHF Omnidirectional Radio Range (VOR) system at Patuxent River Naval Air Station, Md., Philipsburg, Pa., and Ogden, Utah.

The method of employing shoran, which was used for position determination, will be outlined, indicating the techniques of data collection and reduction which were used.

The component errors in the VOR system will be analyzed, including the ground station, airborne receiver, and test equipment accuracies. The effect of aircraft heading and attitude, and of terrain on system accuracy, will be explored.

144. DYNAMIC ASPECT OF ERRORS IN RADIO NAVIGATION SYSTEMS, PARTICULARLY IN CASE OF FAST-MOVING RECEIVERS AND TRANSMITTERS

H. Busignies
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

Under conditions encountered on the earth's surface and in the atmosphere and stratosphere, radio-navigation signals observed on board a moving vehicle show variations. The frequency spectrum and amplitude of such variations can be predicted. They are dependent upon number and position of reflections, relative speed, carrier and modulation frequencies, and polarization. Similar effects are observed when the receiver is fixed and the transmitter is moving.

The present and future speed of aircraft require that these effects be considered in the selection of parameters of radio-navigation systems.

Integrating circuits and spectrum transmissions can be utilized to reduce the dynamic errors. Examples are given which are applicable to instrument landing systems, omnidirectional ranges, DME, direction finders, and generally to all radio aids to navigation.

145. A NEW BASIS FOR ANALYZING RADIO NAVIGATION AND DETECTION SYSTEMS

N. L. Harvey
(Sylvania Electric Products Inc., Bayside, L. I., N. Y.)

The fundamental requirements imposed on radio navigation and echo ranging systems under interference conditions are considered from the viewpoints of the transmission of information and of signal correlation. It is shown that transmission bandwidth and receiver bandwidth may be treated as entirely independent system parameters with the receiver bandwidth used in determining S/N ratio, power, and information capacity; and transmission bandwidth in determining the resolution. The correlation function is used to establish the relationship between the resolution characteristic and the power density distribution of the signal spectrum, regardless of how the spectrum is generated.

146. STOCHASTIC PROCESSES AS APPLIED TO AERIAL NAVIGATION AND DIRECTION FINDERS

L. A. de Rosa
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

In radio navigation, the sources of "determinate" and "random" errors are discussed with means for their reduction. Noise can be considered as a random error, and statistical processes may be used to obtain readings under severe noise conditions. It is shown that the use of auto- and cross-correlation techniques, may, in certain cases enable the ready exchange of reading time, accuracy, and signal-to-noise ratio. Prediction can be used to increase further the available information for exchange. Representative applications of the principles to direction-finding and omnidirectional radio ranges are presented

147. 1,000-Mc CRYSTAL-CONTROLLED AIRBORNE TRANSMITTER FOR DISTANCE MEASURING EQUIPMENT

B. Warriner
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

An airborne pulsed transmitting oscillator for service in Distance Measuring Equipment is described. This oscillator uses a 2C39 tube in a Colpitts circuit with a half-wave coaxial cavity. A frequency range of 965 to 1,088 Mc with a peak-power output of 1 kw was used in a 50-channel system, but in the current RTCA system only ten channels in the 960- to 985-Mc band are used. It is designed to reach 40,000 feet altitude without pressurization and to have extreme tuning linearity to allow stable operation of the A.F.C. servo system.

An ac motor controlled by a miniature double triode is used for frequency control. A 45-Mc quartz crystal is multiplied to about 500 Mc and the multiplier second harmonic is heterodyned against the transmitter and fed into a 60-Mc discriminator. The output of the discriminator controls the motor. Transmitter frequency is changed merely by switching quartz crystals.

SYMPOSIUM
Sound Recording

148. NOISE CONSIDERATIONS IN AUDIO SYSTEMS

F. L. Hopper
(Western Electric Co., Hollywood, Calif.)

Noise limitations of various types of sound recording media are discussed. With improvement in inherent volume range in such a medium as magnetic recording, noise limitations imposed by the audio system require consideration. Noise may be internally generated in the system or may be introduced from extraneous sources by electromagnetic coupling of circuit exposures to interfering fields. Radio- and audio-frequency disturbances, cross talk, thermal noise, shot effect, microphonics, ac hum, and switching transients are some of the forms of disturbance considered.

149. CONSIDERATIONS OF NOISE IN SOUND RECORDING AND REPRODUCING SYSTEMS

Albert W. Friend
(Radio Corporation of America, Princeton, N. J.)

The definitions and measuring techniques related to noise in sound recording and reproducing systems which have been most recently suggested by the IRE Committee on Sound Recording and Reproducing are discussed, with particular emphasis on their application to magnetic record systems. System, equipment, and medium noise and their measurement are discussed separately. The consideration of noise due to the recording medium is further subdivided into residual and modulation noise.

150. MAGNETIC RECORDING FREQUENCY RESPONSE—MEASUREMENT PROCEDURES AND PITFALLS

R. E. Zenner
(Armour Research Foundation, Chicago, Ill.)

A discussion of frequency-response measurements in magnetic recording, including definitions, recommended procedures, and description of difficulties encountered in such measurements.

151. DISTORTIONS IN RECORDING SYSTEMS

H. E. Roys
(Radio Corporation of America, Camden, N. J.)

This paper will describe the distortions encountered in recording systems and discuss problems encountered in measuring them.

152. PERCEPTIBILITY OF FLUTTER IN RECORDED SPEECH AND MUSIC

Harry Schecter
(Massachusetts Institute of Technology, Cambridge, Mass.)

A form of distortion common to sound records of all types is flutter or "wow" caused by irregularities in speed of the recording medium. These speed irregularities produce frequency modulation of the recorded signals. To establish realistic tolerances for flutter, we need quantitative subjective data on the relative perceptibility of flutter as a function of the extent and rate of modulation for typical program material under typical listening conditions. The problem of obtaining such data will be outlined and some preliminary results will be presented.

Propagation II
Impact of Propagation on Operation of Systems

153. A MICROWAVE PROPAGATION TEST

J. Z. Millar and L. A. Byam, Jr.
(Western Union Telegraph Co., New York, N. Y.)

This paper describes a microwave propagation test which was conducted over a period of a year with simultaneous transmission on wavelengths of 16.2, 7.2, 4.7 and 3.1 cm over an unobstructed 42-mile overland path. Comparative charts depict variations in daily fading range, illustrate diurnal and seasonal influences on fading and reveal the marked disparagement between winter and summer fading. Of particular significance, curves are offered showing relative field-strength distribution for both winter and summer periods. Additional curves are included relating to distribution of hourly minima with respect to four arbitrarily selected signal levels. The various curves are useful in considerations bearing on continuity of service that may be expected with relation to wavelength and to time of day, winter or summer. Pictures of the propagation tower, transmitter units, control and recording equipment are included.

154. DIVERSITY RECEPTION TECHNIQUES

S. H. Van Wambeck
(Washington University, St. Louis, Mo.)
A. H. Ross
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

An outline is given of the work on diversity reception techniques being carried out jointly by the Signal Corps and Washington University, the purpose of which is to determine the characteristics, limitations, and relative merits of space, polarization, and frequency diversity systems.

Tests conducted over a 900-mile circuit are described, and results obtained with various antenna spacings and configurations are shown in simple graphic form. The results can be expressed in terms of improvement over the performance of a simple reference doublet antenna. Application of the engineering information derived from this investigation to military radio communication problems is explained.

155. EXPERIMENTAL EVALUATION OF DIVERSITY RECEIVING SYSTEMS

John L. Glaser and S. H. Van Wambeck
(Washington University, St. Louis, Mo.)

Methods used and results obtained in a long-range experimental study of fading with ordinary and diversity receiving sys-

tems are discussed. Results are expressed statistically in terms of the nonusable circuit time, and also in terms of the number of fades per minute below least usable level.

Data for this analysis are obtained on semi-automatic equipment which measures the total time the signal spends in each of seven pre-established intervals of signal strength and counts the number of times the signal enters each interval. The instruments thus accomplish at the time of recording a substantial part of the analysis.

Tests which have been conducted continuously over a period of about twenty months at frequencies of 7, 12, and 16 Mc have included polarization, and dual and triple spaced-antenna diversity systems. Typical results are presented.

156. COMPARISON OF MODULATION METHODS FOR VOICE COMMUNICATION OVER IONOSPHERIC RADIO CIRCUITS

M. G. Crosby
(Crosby Laboratories, Mineola, L. I., N.Y.)
H. F. Meyer and A. H. Ross
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

Intelligibility tests conducted over an ionospheric radio circuit and through a multipath simulator using double-sideband amplitude modulation, narrow-band frequency modulation, and narrow-band phase modulation with various demodulation systems are described. The use of the simulator to verify mathematical analysis of the results obtained in evaluating the effects of multipath transmission are explained. The effectiveness of exalted carrier reception in raising the average intelligibility for amplitude or phase modulation in the presence of multipath transmission is demonstrated and the modifying effects of noise and interference on the relative merits of the various systems are discussed.

157. COMPARISON OF MODULATION METHODS FOR FACSIMILE COMMUNICATION OVER IONOSPHERIC RADIO CIRCUITS

M. Acker and B. Goldberg
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

A description is given of intelligibility tests conducted on both an actual ionospheric circuit and on a multipath transmission simulator to evaluate the performance of various methods of transmitting facsimile signals over a radio circuit under multipath transmission conditions. Subcarrier FM on AM and on narrow-band PM and carrier shift modulation systems were used to transmit the intelligence together with various demodulation methods. Results obtained from the tests conducted over the actual ionospheric circuit are analyzed by statistical methods and the results presented in the form of intelligibility ratings. A discussion of the effects of noise, interference, and multipath transmission on the various methods is given. The results of the tests conducted through the multipath transmission simulator are presented in the form of a novel three-dimensional intelligibility plot. Theoretical considerations and correlation between the laboratory and actual ionospheric circuit tests are discussed.

Electron Tubes IV

Materials and Techniques

158. A VACUUM SEAL BETWEEN METALS AND CERAMICS FOR HIGH TEMPERATURE APPLICATIONS

H. W. Soderstrom and K. H. McPhee
(Collins Radio Co., Cedar Rapids, Iowa)

A new method for joining metals to insulators called the "Somac" is described. The materials involved are not required to have matched coefficients of expansion. For most practical purposes there are no size limitations. The operating and outgassing temperatures can be as high as 1100° C. The short overlap of metal and porcelain in this seal makes it essentially a butt seal. Precision alignment in assembly is possible. Other applications will be discussed.

159. EFFECT OF COATING COMPOSITION OF OXIDE-COATED CATHODES ON ELECTRON EMISSION

E. G. Widell and R. A. Hellar
(Radio Corporation of America, Harrison, N. J.)

Investigation of oxide-coated cathodes has been made to determine the effect on electron emission of varying the proportions of emitting oxides. The results show that maximum electron emission under saturation conditions is obtained from a solid solution containing strontium oxide (SrO) and barium oxide (BaO) in an approximate molecular ratio of 7 to 3. Increased emission is obtained by the further addition of calcium oxide to the barium and strontium oxides.

These results also indicate that maximum size of the coprecipitated barium and strontium carbonate particle occurs at the same molecular ratio as that giving maximum electron emission from the oxides.

When the saturation emission obtained with a square-wave voltage pulse was measured, the current pulse as observed on a synchroscope was also square wave and showed no measurable decay characteristics for a pulse duration of 10 seconds.

160. EFFECTS OF CONTROLLED IMPURITIES IN NICKEL CORE METAL ON THERMIONIC EMISSION FROM OXIDE-COATED CATHODES

George Hees
(Sylvania Electric Products Inc., Kew Gardens, L. I., N. Y.)

Emission measurements were made over a 500-hour life period on oxide coated cathodes having 18 different binary nickel-alloy core metals. The samples of nickel and of the additive (both of highest purity) were vacuum melted together in order to provide a rigid control on the composition of each alloy. All tubes were processed under the same controlled conditions.

An empirical emission equation

$$I = A' \exp\left(-\frac{e\psi}{kT}\right)$$

was used. The parameters ψ and A' were found to vary in such a manner that, of life, the following relationship existed:

$$\log A' = k\psi + B$$

where k and B are constants.

Some alloys which appear to be especially suitable as base metals for oxide-coated cathodes are aluminum-nickel, chromium-nickel, molybdenum-nickel, and thorium-nickel. Alloys which resulted in poorest emission were boron-nickel, beryllium-nickel, and iron-nickel.

161. INVESTIGATION OF CONTAMINANT IN VACUUM TUBES

Paul D. Williams
(Eitel-McCullough, Inc., San Bruno, Calif.)

In an ONR sponsored program, tube materials are being examined on a mass spectrometer to determine the volatile components released in vacuum during heating. Of four glasses normally used in tube construction, nonex glass continued to evolve water after extended bake-out period.

Gases evolved from the blank and tube elements lose their identity by reaction with other elements. Volatile component of heated Al-200 ceramic, while injuring carburized thoriated tungsten emission does not destroy emission. Volatiles from quartz heated to softening temperature does not injure carburized thoriated tungsten emission. Volatile components of ceramic A-243 appear to destroy emission.

162. HOT STRENGTH PROPERTIES OF FILAMENTARY ALLOYS

Bernard Wolk
(Sylvania Electric Products Inc., Kew Gardens, L. I., N. Y.)

The results of an investigation of the hot strength properties of several nickel- and cobalt-based alloys are given. These materials have as their minor constituents such elements as aluminum, tungsten, chromium, etc., and were chosen for their possibilities as efficient primary electron emitters (when coated with alkaline earth oxides) over at least 500 hours of life.

The general term hot strength is defined as both resistance to flow, and time to rupture at temperatures around 800°C. A stress range was chosen so that accelerated conditions prevail, and a measure of the relative strength of a variety of filamentary alloys is thereby practicable.

Components

163. MINIATURIZATION TECHNIQUES: A DISCUSSION AND PROPOSAL

M. Abramson and S. Danko
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

Objective analysis of hand-wiring, printed circuit, and composite systems of miniature circuit fabrication show inherent advantages and weaknesses in each system. The best features of these approaches have been combined into a proposed composite system which promises lower costs and higher production through assembly simplification. This proposed approach, designated as

"Auto-Sembly" requires no elaborate or expensive tooling and no unfamiliar skills. The "Auto-Sembly" approach is based on the use of low cost prefabricated conductor patterns on an insulated surface, separately fabricated components, rapid components-to-pattern assembly and final protective packaging.

164. THE EXPONENTIAL-LINE PULSE TRANSFORMER

E. R. SCHATZ AND E. M. WILLIAMS
(Carnegie Institute of Technology, Pittsburgh, Pa.)

This paper describes the theory and design of exponential-line section transformers for very short rise time and extremely high-power short pulse requirements.

A general treatment of the theory of transients in transmission lines with exponential taper and negligible losses is given, and an approximate treatment for the effect of losses is included. The problems of choice of parameters and suitable dielectrics for economical transformer design are considered. Two typical designs are described; (1) an experimental one to two, 2-megawatt, 0.04-microsecond transformer together with experimental results and (2) the design of a proposed 15- to 100-kv., 2.03-microsecond, 200 megawatt transformer to be used in an electric ion deflector in a proton synchrocyclotron.

165. UNIVERSAL PRECISION RESOLVERS

DONALD L. HERR
(Reeves Instrument Corp., New York, N. Y.)

The basic equations and equivalent circuits of idealized and physical high-precision, universal-frequency, ac electromagnetic resolvers are presented. The influences upon optimum performance and design of the numbers and skewnesses of stator and rotor poles, of stator and rotor winding distributions and quality, of magnetic material, of physical dimensions, and of manufacturing processes are given. The design requirements for high-precision resolution with a maximum functional error of ± 0.05 of 1 per cent over 360° of shaft rotation, at all three common carrier-frequencies of 60, 400 and 1,000 cps, with the same resolver, are presented. Particular attention is given to space-harmonic suppression by the resolver itself, and time-harmonic and quadrature noise suppression by the resolver and compensating—booster-amplifier combination in a minimum weight and volume design now in mass production.

166. A COMPACT MAGNETIC MEMORY

PAUL L. MORTON
(University of California, Berkeley, Calif.)

A drum of standard aluminum tubing 8¼ inches in diameter and 27 inches long, coated with magnetite and driven at 3,600 rpm by a ¼ horsepower motor, has been developed to store 10,000 ten-decimal-digit numbers at a cell density of over 900 binary digits per square inch. Combination reading-recording heads operate at a frequency of 144 kc, with writing currents of about 150 milliamperes and voltages when reading of about 0.1 volt. Tests have shown these noncontact heads can be operated up to 300 kc and 3,000 cells per square inch.

167. SYNCHRO-CYCLOTRON FIELD REGULATOR

C. S. MCKOWN
(Sperry Gyroscope Company, Great Neck, L. I., N. Y.)
WILLIAM P. CAYWOOD, JR.
(Carnegie Institute of Technology, Pittsburgh, Pa.)

High-energy nuclear-particle accelerators present many design problems of a new type and magnitude. Such a problem is the design of a regulator to control and maintain the field of the Carnegie Institute of Technology 425 mev synchro-cyclotron to less than 1 gauss in 20,000.

This paper contains a description of the salient features of a proposed design, "Synchro-Cyclotron Progress Report," October 1, 1949; Navy contract N7onr-303, Task Order 1.) in particular: (a) a method of applying a proton-controlled oscillator as a measuring device for the magnetic field, and (b) a mathematical analysis of the electrical loop which includes the cyclotron field system with a time constant of one minute. The means of obtaining very rapid response in spite of there being three different circuits having significantly large time constants (two cascaded excitors in addition to the synchrocyclotron field) is considered of special interest.

Oscillators

168. THE TRANSIENT BEHAVIOR OF A CLASS-C OSCILLATOR

CHESTER H. PAGE
(National Bureau of Standards, Washington, D. C.)

When a stable oscillator is switched on, the bias and amplitude approach their equilibrium point via a spiral path in the bias-amplitude phase plane.

Two types of instability exist: (1) the common squegg or essentially intermittent oscillation, and (2) the weak squegg, wherein the bias and amplitude undergo small periodic variations.

The weak squegg is characterized by a small elliptical limit cycle replacing the stable point. The strong squegg exhibits a large sharp-pointed limit cycle, which surrounds either a weak squegg cycle or a stable point, providing two allowable states of operation. In the latter case, the oscillator will always squegg when turned on, but can be forced into the stable state.

169. MODE SUPPRESSION IN BROAD-BAND REFLEX KLYSTRON OSCILLATORS

A. H. SONNENSCHEIN AND H. A. FINKE
(Polytechnic Research and Development Co., Inc., Brooklyn, N. Y.)

The design of resonators for externally tuned reflex klystron oscillators is greatly complicated by the problem of mode interferences at frequencies other than the desired. The existence of multifrequency bunching is an additional complication. These problems can be solved by increasing the resonator losses to the undesired modes beyond the threshold required to sustain oscillation. This is accomplished by means of microwave filters, so as not to affect the desired mode. The filters described result in very simple mechanical structures, requiring neither tuning nor tracking, as the oscillator is tuned over its desired range. The practical application of these methods will be demonstrated by describing two oscillators which have been designed and built. They cover without hiatus, the band from 3,600 to 11,000 Mc per second, using the same tube.

170. TELEMETERING BLOCKING OSCILLATOR

W. TODD
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

The theory of blocking oscillators for telemetering purposes is derived along with an approximate equivalent circuit and formulas. The choice of proportions of time constants and the mode of operation are discussed from the standpoint of absolute frequency stability, non-proportional drift, pulse width, power dissipation, and tube characteristics. Some design-difficulties of oscillators of this type and their corrective means are discussed. Conclusions are drawn as to the most satisfactory design of such an oscillator for a typical radiosonde telemetering application.

171. SOME ASPECTS OF RF PHASE CONTROL IN MICROWAVE OSCILLATORS

E. E. DAVID, JR.
(Massachusetts Institute of Technology, Cambridge, Mass.)

The effect of an injected rf signal on a free-running oscillator may be described exactly by means of a graphical construction on the Rieke Diagram. The analysis may be generalized to include mutual coupling of two or more oscillators. The constant phase predicted is not observed in practice, the discrepancy being due to residual hum on the oscillator electrodes. By utilizing this effect, either phase or amplitude modulation may be obtained.

When the oscillator to be synchronized is pulsed, there are present (attendant to the starting disturbance) certain transient conditions which cannot be described properly by the steady-state theory. In particular, the phase transient persists several times longer than the rf voltage build up. Therefore, the preoscillation noise-to-locking signal ratio is an important parameter in determining pulse-to pulse coherence.

172. SEVEN-LEAGUE OSCILLATOR

F. B. ANDERSON
(Bell Telephone Laboratories, Inc. New York, N. Y.)

This is a wide-range variable frequency bridge-type oscillator which, in a preliminary form, has been demonstrated to be tunable over a range of 20 cps to 3 Mc in one sweep of a linear control. The frequency scale is approximately logarithmic. Accuracy of the order of ± 0.5 per cent is attainable with ordinary parts. Frequency stability is of the order of ± 2 per cent per db of the tube gain variation in the lower decades, but falls off at high frequencies. A frequency range of one billion to one extreme ratio appears to be possible.

Abstracts and References

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ACOUSTICS AND AUDIO FREQUENCIES

016:534 1

References to Contemporary Papers on Acoustics—A. Taber Jones. (*Jour. Acous. Soc. Amer.*, vol. 21, pp. 556–563; September, 1949.) Continuation of 2995 of 1949.

534.213.4 2

On the Propagation of Sound Waves in Narrow Conduits—O. K. Mawardi. (*Jour. Acous. Soc. Amer.*, vol. 21, pp. 482–486; September, 1949.) An approximate solution of sufficient accuracy for narrow tubes of arbitrary shapes is derived and applied to the case of a wire-filled tube. Losses due to viscosity, radiation, and thermal conduction are taken into account. Theoretical predictions are in satisfactory agreement with experimental results.

534.232 3

Theory of Focusing Radiators—H. T. O'Neil. (*Jour. Acous. Soc. Amer.*, vol. 21, pp. 516–526; September, 1949.) The sound field due to a concave spherical radiator vibrating with uniform normal velocity is determined approximately. The radius of the circular boundary of the radiator is assumed to be large compared with λ and with the depth h of the concave surface. The point of greatest intensity is not at the center of curvature but approaches it with increasing values of h/λ. The greatest intensity is not much greater than the intensity at the center of curvature except when h/λ is small. The calculations are in reasonable agreement with Willard's experimental data (3003 of 1949). See also 8 below.

534.232 4

The Theory of Sound Vibrations in Open Tubes—L. A. Weinstein. (*Zh. Tekh. Fiz.*, vol. 19, pp. 911–930; August, 1949. In Russian) The radiation of a symmetrical sound wave from the open end of a tube is discussed and a rigorous solution of the problem is given. The results obtained are also applied to the case of asymmetrical waves.

534.25 5

Refracting Sound Waves—W. E. Kock and F. K. Harvey. (*Jour. Acous. Soc. Amer.*, vol. 21, pp. 471–481; September, 1949.) Refracting structures consisting of arrays of obstacles which are small compared with λ are described. These obstacles increase the effective density of the medium and thus reduce the velocity of sound passing through the array. When the dimensions of individual obstacles approach $\lambda/2$, the effective refractive index varies with λ and prisms then cause a dispersion of the waves. Path-length delay-type lenses for focusing sound waves are also described. A diverging lens is discussed which can be used in conjunction with an exponential horn loudspeaker to produce an approximately circular wave front at the higher audio frequencies. For em lenses similar in principle see 2176 of 1948 (Kock). See also *Bell Lab. Rec.*, vol. 27, pp. 349–354; October, 1949.

534.321.9 6

Ultrasonics in Fluids—E. G. Richardson. (*Nature* (London), vol. 164, pp. 772–773; November 5, 1949.) Report of a British Association symposium. Most of the papers deal with applications of the ultrasonic interferometer.

534.321.9:534.833.4 7

An Optical Method for the Determination of Ultrasonic Absorption in Opaque Soft Media—T. Hüter and R. Pohlman. (*Z. Angew. Phys.*, vol. 1, pp. 405–411; June, 1949.) Two methods are described which are particularly suitable for materials with low acoustic impedance and high absorption. The methods depend on the diffraction of light by ultrasonic waves and on the brightness distribution in the various orders of the diffraction spectrum. Results are given for various animal tissues and also for a few artificial insulating materials.

534.321.9:549.514.51 8

Ultrasonic Radiation from Curved Quartz Crystals—L. Fein. (*Jour. Acous. Soc. Amer.*, vol. 21, pp. 511–516; September, 1949.) The electroacoustic efficiencies at 1250 kc of four one-inch square x-cut quartz crystals are deduced from transmitting frequency responses in water, admittance measurements in air and water, and radiation patterns in water. One crystal is plane; the others have radii of curvature of 25, 7, and 4 cm, respectively; their thicknesses are practically equal. Efficiency values deduced from acoustic measurements do not agree as well with calculated potential efficiencies as do those derived from admittance measurements; the latter indicate that the radiation resistance is the same for the 4 crystals. The point of maximum acoustic intensity is not necessarily at the center of curvature of the crystal. The admittance measurements in air indicate that the effective mass of a crystal decreases with increasing curvature. See also 3 above.

534.41:534.78:621.385.832 9

The Cathode-Ray Sound Spectroscope—R. C. Mathes, A. C. Norwine, and K. H. Davis. (*Jour. Acous. Soc. Amer.*, vol. 21, pp. 527–537; September, 1949.) A device for the rapid analysis of short samples of speech and other sounds. The energy/frequency distribution of the sound at a particular instant is displayed as a two-dimensional pattern, and the distribution over an interval as a three-dimensional pattern. The sample is recorded slowly on a magnetic disk, played back 200 times as fast, and analyzed by a broad-band high-frequency system.

534.612.4 10

On the Reciprocity Free-Field Calibration of Microphones—W. Wathen-Dunn. (*Jour. Acous. Soc. Amer.*, vol. 21, pp. 542–546; September, 1949.)

534.75 11

Some Determinants of Interaural Phase Effects—I. J. Hirsh and F. A. Webster. (*Jour. Acous. Soc. Amer.*, vol. 21, pp. 496–501; September, 1949.) The binaural threshold for a pure tone presented against a background of noise depends on the interaural phase differences of the tone and the noise. Results of experiments in which a 250-cps tone was presented against four different kinds of noise background indicate that the threshold for the tone in the presence of a periodic masking sound is not significantly dependent on interaural phase relations, but the masking and interaural phase effects increase as the frequency band of a random masking sound approaches the frequency of the tone.

534.78 12

Extraction and Portrayal of Pitch of Speech Sounds—O. O. Gruenz, Jr., and L. O. Schott. (*Jour. Acous. Soc. Amer.*, vol. 21, pp. 487–495; September, 1949.) An improved method using a combination of gain control, double detection, voiced sound selection, unvoiced sound exclusion, and a means of counting the fundamental vibrations in the voiced sound intervals. Reliable indications of pitch have been obtained for frequencies in the range 100 to 600 cps for a wide variety of voices. The pitch-indicating signals have been applied to a number of visual portrayal devices such as that noted in 3520 of 1946 (Riesz and Schott).

534.861:534.76 13
Stereophony—K. de Boer. (*Tijdschr. ned. Radiogenoot.*, vol. 14, pp. 137–146; September, 1949.) General discussion of the basic principles of recording and of reproduction for an audience.

621.392.51 14
Theoretical Aspects of the Reciprocity Calibration of Electromechanical Transducers—S. P. Thompson. (*Jour. Acous. Soc. Amer.*, vol. 21, pp. 538–542; September, 1949.) A unified theory is presented in a form from which particular calibration theories can be derived. Two types of calibration procedure are discussed as special cases.

621.395.61 15
Directional Microphone—H. F. Olson and J. Preston. (*RCA Rev.*, vol. 10, pp. 339–347; September, 1949.) The development of a second-order gradient system, with ribbon-type units, is described. The system has a uniform and narrow directivity pattern, and a smooth frequency-response characteristic between 50 and 15,000 cps. In conventional studios, speech can be picked up at a distance of 12 feet. By using a monitoring console in conjunction with several of these microphones in fixed positions, so that each covers a section of the studio area, large and rapidly changing areas of action can be covered with less variation in output level than is possible with the conventional microphone and boom arrangement.

621.395.623.7 16
Third-Class Loudspeakers with Low Power Consumption—P. V. Anan'ev. (*Radiotekhnika* (Moscow), vol. 4, pp. 18–27; July and August, 1949. In Russian.) Cheap loudspeakers, which would be effective in operation and whose construction would not require materials in short supply, are needed for remote collective farms. Loudspeakers using Rochelle salt crystals appear to meet the requirements. Such loudspeakers, provided with a volume control and a special correcting circuit, are described in detail; many performance curves are included.

621.395.623.7 17
New 15-Inch Duo-Cone Loudspeaker—H. F. Olson, J. Preston, and D. H. Cunningham. (*Audio Eng.*, vol. 33, pp. 20–22, 48; October, 1949.) Modifications of the RCA loudspeaker Type LC1A which make mass production possible are discussed. See also 993 or 2664 of 1947 (Olson and Preston).

621.395.623.8 18
Ground Loudspeakers—D. Scott. (*Audio Eng.*, vol. 33, pp. 18–19; October, 1949.) *See* 19 below.

621.395.623.8 19
Underground Loudspeakers—J. Merhaut. (*Tesla Tech. Rep.* (Prague), pp. 35–38; March, 1949.) Description of the loudspeakers set level with the ground in the Strahof stadium at Prague for use in connection with mass gymnastic and other displays in an arena 1,000 feet by 650 feet. The diaphragm is made of synthetic humidity-resisting material. The horn is of the folded exponential type. The complete unit is sealed and thus protected against humidity. Performance is unaffected even by rain water. Arrangements are provided for easy maintenance or replacement of the loudspeaker unit without removal of the main cover, which is strong enough for a truck to be driven over it.

621.395.625.2 20
Determination of the [stylus] Velocity Amplitude in "Constant-Velocity" Recording, by Measurement of the Width of a Luminous Band—J. H. (*Radio Tech. Dig.* (Franç), vol. 3, pp. 303–305; October, 1949.) Short description of the method of Buchmann and Meyer, in which the disk groove is illuminated by a parallel beam of light and the pattern formed by the reflected beam on a screen a few meters away is examined.

621.395.625.3 21
New Type of Magnetic Recorder—L. A. Fishoff. (*Radio Tech. Dig.* (Franç), vol. 3, pp. 291–294; October, 1949.) Short description of the principal features of a tape recorder in which tape friction is avoided by using magnetic coupling. Some details of the mechanism, the amplifier, and the motors are included.

621.395.625.3 22
Magnetic Recording—M. Alixant. (*Radio Tech. Dig.* (Franç.), vol. 3, pp. 259–291; October, 1949. Bibliography, pp. 294–301.) Basic principles and modern tapes and recording heads are discussed. Details are given of many commercial instruments for recording on wire or tape. Most of these are of American manufacture, but some French, Swiss, German, and English recorders are included.

621.395.625.3:621.395.813 23
Techniques for Improved Magnetic Recording—L. C. Holmes. (*Elec. Eng.*, vol. 68, pp. 836–841; October, 1949.) Essential substance of a 1948 National Electronics Conference paper. Summary noted in 2423 of 1949.

621.395.92 24
A Modern Hearing-Aid—(*Radio Tech.* (Vienna), vol. 25, pp. 613–615, 622; October, 1949.) Description, with detailed circuit diagram, of the "Vienna" lightweight equipment using subminiature components. The frequency-response curve can be adjusted and agc is provided. The crystal microphone, amplifier, and batteries are enclosed in a flat molded case, 12×6.5×2.5 cm, which can easily be carried inside a coat or in a pocket. Four models are available, suitable for different degrees of deafness. One model is of a simpler type without tone adjustment.

621.395.625 25
Elements of Sound Recording [Book Review]—J. G. Frayne and H. Wolfe. Publishers: J. Wiley and Sons, New York, 1949, 686 pp., $8.50. (*Electronics*, vol. 22, pp. 233–234; November, 1949.) The first 12 chapters review fundamental material on microphones, amplifiers, attenuators, equalizers, etc. The main interest of the authors is in sound film recording. "... a very welcome addition to the literature of sound recording."

ANTENNAS AND TRANSMISSION LINES

621.315.2 26
Highly Balanced Radio-Frequency Transmission Lines—K. H. Zimmermann. (*Elec. Commun.*, vol. 26, pp. 201–203; September, 1949.) The twinax line differs from a coaxial line in having two identical inner conductors which are insulated separately and then twisted. The same dielectric material is extruded over the inner conductors after twisting. The braided shield and thermoplastic jacket are then applied as for a coaxial line. Formulas and graphs are given for the characteristic impedance of twinax lines; the practical determination of various parameters is briefly discussed.

621.315.212 27
Coaxial Cable with Confocal Elliptical Cylindrical Conductors—M. R. Shebes. (*Radiotekhnika* (Moscow), vol. 4, pp. 36–44; July and August, 1949. In Russian.) The effects of deformation of the cylindrical conductors of a coaxial cable on its parameters are discussed. The results obtained for the characteristic impedance and attenuation are compared with those obtained for cables with (a) non-concentric circular cylindrical conductors, (b) cylindrical conductors whose cross-sections are bounded by limaçons.

621.392.1 28
Terminal Impedance and Generalized Two-Wire-Line Theory—R. King and K. Tomiyasu. (Proc. I.R.E., vol. 37, pp. 1134–1139; October, 1949.) Conventional transmission-line theory neglects variations in the parameters of the line, and coupling near a terminating impedor or near any other departure from uniformity. A theory is derived in which a simple terminal-zone network N of lumped series and shunt elements is introduced to allow for these effects. Such a network can be determined for each type of termination or discontinuity. The apparent terminal impedance z_{sa} which is the impedance actually measured on a lossless line at a distance $\lambda/2$ from the termination, is the impedance of the network formed by combining N with the theoretical isolated impedance Z_s of the load. For a fixed termination and given Z_s, Z_{sa} may vary greatly with the nature of the connection to the line, the relative orientation of line and load, and the type of line and its dimensions.

621.392.1 29
Energy Relationships for a High Frequency Transmission Line—A. L. Fel'dshteyn. (*Radiotekhnika* (Moscow), vol. 4, pp. 45–50; July and August, 1949. In Russian.) Discussion taking account of losses and specified terminations. Conditions for optimum energy transfer are established and a conception of Q for the combination of oscillator, line, and load is introduced.

621.392.26† 30
Spatial Beating in Coupled Waveguides—P. E. Krasnushkin and R. V. Khokhlov. (*Zh. Tekh. Fiz.*, vol. 19, pp. 931–942; August, 1949. In Russian.) Spatial beating can be observed in two parallel coupled waveguides in each of which symmetrical waves are propagated. The phenomenon consists of a periodic transference of wave energy from one waveguide to the other. The phenomenon can be regarded as an analogy in space of the beating which takes place between two coupled oscillatory systems.

The case of two semi-elliptical waveguides coupled through a slot is here investigated theoretically and experimental results are given.

621.392.26†:517.54 31
Application of Conformal Representation to the Field Equations for Rectangular Waveguides of Non-Uniform Cross-Section—R. Piloty, Jr. (*Z. Angew. Phys.*, vol. 1, pp. 441–448; August, 1949.) By means of conformal transformation, the field at an irregularity in either the E-plane or the H-plane of a rectangular waveguide can be referred to the field in a plane-walled guide filled with a medium whose permeability or dielectric constant varies from point to point. For excitation by $H_{1,0}$ waves, a relatively simple partial differential equation for the field can be derived. It is also shown that by means of a simple frequency transformation the field at an E-irregularity in a rectangular waveguide can be referred to that at a corresponding irregularity in a band-pass transmission line. A method of solving the general equation, which among other things enables the impedance transformation of the waveguide quadripole to be determined, will be given in a later paper. See also 2427 and 2428 of 1949 (Rice).

621.392.26†:621.392.5 32
Basis of the Application of Network Equations to Waveguide Problems—Kerns. (*See* 56.)

621.392.3.012.1 33
A Simple Vector Diagram for High-Frequency Lines—P. Cornelius. (*Commun. News*, vol. 10, pp. 33–40; June, 1949.) The input impedance of a lossless hf transmission line can be determined, for a given terminating impedance, by means of a Smith chart; the attenuation is only given if the line is terminated by its characteristic impedance. The impedance of a line with losses and terminated arbitrarily can be determined by the vector method here given, but the attenuation is only given for termination by the characteristic impedance. The vec-

tor method is also used to derive various well-known relations.

621.392.43 34
Two-Band Antenna-Matching Networks—J. G. Marshall. (*QST*, vol. 33, pp. 14–18 and 48–51, 114; October and November, 1949.) Continuation of 3869 of 1945. All practical cases of antennas working on two harmonically related amateur bands are discussed, with straightforward design formulas.

621.396.67 35
Short-Antenna Characteristics—Theoretical—L. C. Smeby. (Proc. I.R.E., vol. 37, pp. 1185–1194; October, 1949.) A mathematical analysis of experimental data obtained by Smith and Johnson (651 of 1948) for antennas with umbrella top-loading. The theory developed agrees satisfactorily with the experimental results. For single-tower operation, the horizontally polarized radiation is negligible. With the optimum length of umbrella, the vertical radiation characteristic is the same as it would be from the radiator without top loading. The method of analysis can be extended to other kinds of top loading.

621.396.67:621.397 6 36
"Supergain" TV Antenna Developed by RCA—(*Broadcast News*, no. 56, pp. 4–6; September, 1949). An antenna consisting of dipoles mounted in front of reflecting screens $\frac{1}{2}\lambda$ wide and $\frac{5}{8}\lambda$ high fixed on the four sides of a tower. The screens are connected electrically to the tower and to each other at their vertical edges. Various methods of connection to obtain directional patterns of different shapes are possible; rectangular dipole shielding wings connected to the edges of the screens at an angle of 135° can be used to prevent undesired mutual coupling of the dipoles. The antenna is usually fed by a single transmission line. Radiators are spaced 0.9λ apart vertically, so that current distribution is nearly uniform. The power gain relative to a dipole, averaged over 360° azimuth, is $1.1n$, where n is the number of bays of dipoles.

621.396.67:621.397.6 37
A Tunable Built-In TV Antenna—R. B. Albright. (*Electronics*, vol. 22, pp. 134..150; November, 1949.) This Philco antenna is a $\lambda/2$ dipole consisting of two tapered sections of Al foil 0.005 inch thick. Signals can be received efficiently from all the 12 existing U. S. vhf television channels. A tunable matching circuit, which includes a variable capacitor and three fixed inductors, enables the user to tune in to each channel so as to obtain the best response and to eliminate interference. The equivalent circuit of the complete system and resonance conditions are discussed for the high and the low television frequency bands. See also *Tele-Tech*, vol. 8, pp. 37, 60; October, 1940.

621.396.67.029.54 38
Mode of Action and Design of Modern Broadcasting Transmitting Aerials for the Medium-Wave Range—K. Fischer. (*Elektrotech. u. Maschinenb.*, vol. 66, pp. 237–242; September, 1949.) Discussion of ground-wave and space-wave effects, radiation diagrams of ordinary antennas, and special antennas designed to minimize fading.

621.396.671 39
Vertical-Aerial Radiation Characteristics over Uneven Terrain—H. Köhler. (*Elektrotechnik* (Berlin), vol. 2, pp. 297–304; November, 1948.) A detailed theory is given of the directional characteristics of arrays of vertical antennas over uneven ground. A df system described by Stenzel, while giving good results for eliminating the effects of variations of earth constants in the case of level ground, does not give such good results over uneven ground. Measurements of directional characteristics which support the theory are described and the physical connection between measured reflection factors and the electrical properties of the ground is considered.

621.396.677.029.54 40
Design Considerations for Directive Antennae-Arrays at Medium-Wave Broadcast Frequencies, Taking into Account the Final Radio-Frequency Amplifier Circuits—J. C. Nonnekens. (*HF* (Brussels), no. 3, p. 80; 1949. In English.) Correction to 1605 of July.

CIRCUITS AND CIRCUIT ELEMENTS

061.4:[621.317.7+621.38+621.396.69 41
Radiolympia Review.—(*See* 145.)

621.3:512.974 42
Vectorial Space. Region of Representation for Electrotechnics—(*See* 132.)

621.314.3† 43
The Direct-Current Choke—W. Blankenburg. (*Elektrotechnik* (Berlin), vol. 3, pp. 135–140; May, 1949.) Starting from the ordinary premagnetized choke, magnetic amplifier arrangements are developed and their properties are investigated. A null-current model is described with an amplification factor of about 10^7; this is controlled by a thermoelement and gives a load power of about 1 w.

621.314.3† 44
An Experimental Study of the Magnetic Amplifier and the Effects of Supply Frequency on Performance—E. H. Frost-Smith. (*Jour. Brit. I.R.E.*, vol. 9, pp. 358–373; October, 1949.) Magnetic amplifiers have hitherto been used mainly for amplifying small dc powers. The ac supply to the amplifier then operates normally at power frequencies. The time constant of the amplifier is limited by the supply frequency; the possibility of reducing response time by using a supply frequency of about 20 kc is considered. For a given frequency and power output, there is an optimum core size, which becomes more critical as the supply frequency is increased. Experimental results show that the magnetic amplifier has distinct possibilities as an af amplifier.

621.314.3† 45
Magnetic Amplifiers—(*Proc. IEE* (London), part II, vol. 96, pp. 767–768; October, 1949.) Report of an IEE Measurements Section discussion on 2728 of 1949 (Milnes) and 2729 of 1949 (Gale and Atkinson).

621.314.3† 46
The Magnetic Amplifier—P. M. Kintner and G. H. Fett. (*Radio and Televis. News, Radio-Electronic Eng. Supplement*, vol. 13, pp. 14–18..28; August, 1949.) An analysis from first principles, with discussion of factors affecting design.

621.314.3† 47
Magnetic Amplifiers—M. Alixant. (*Radio Tech. Dig.* (Franç), vol. 3, pp. 153–159; June, 1949. Bibliography, pp. 159–163.) Short account of basic principles, practical construction, and applications.

621.316.86 48
Electrolytic Thermistors—F. Gutmann and L. M. Simmons. (*Rev. Sci. Instr.*, vol. 20, pp. 674–675; September, 1949.) A thermistor for use with ac is easily constructed by immersing Pt electrodes in a viscous solution of water glass. A temperature change of only 7.2° at 300°K will double or halve the resistance. This thermistor is three times as sensitive as commercial thermistors, withstands ac potentials of more than 240 v, and its resistance and sensitivity when cold are easily controlled. Other solutions were tested; they could withstand high ac potentials, but were less sensitive.

621.316.86.001.8 49
Properties and Applications of Thermistors—E. Ancel. (*Radio Franç.*, no. 9, pp. 3–10; September, 1949.) Applications to thermometry, temperature compensation and control, power measurement, voltage regulation, agc, and in relays, oscillators, and modulators, are outlined.

621.318.2:621:23 50
Permanent Magnets in Drag Devices and Torque-Transmitting Couplings—R. J. Parker. (*Gen. Elec. Rev.*, vol. 52, pp. 16–20; September, 1949.) Permanent magnets are used in four principal types of torque-transmitting device, namely: (a) eddy-current devices, (b) hysteresis devices, (c) iron/oil magnetic couplings, and (d) salient-pole synchronous couplings. Specimens of each of these types are briefly described, and formulas for the torque are obtained.

621.318.42:621.314.263:621.314.3† 51
The Use of Ferrite-Cored Coils as Converters, Amplifiers and Oscillators—V. D. Landon. (*RCA Rev.*, vol. 10, pp. 387–396; September, 1949.) Theoretical treatment of the behavior of nonlinear inductors used as frequency converters up to a few megacycles shows that (a) if the excitation frequency is higher than the signal frequency and the if, the circuit regenerates; (b) if the excitation is sufficient, the circuit may oscillate at the signal frequency and if simultaneously; (c) if either the rf or the if circuit is tuned to a frequency above the oscillator frequency, the circuit is degenerative. Experimental results and transformer specifications are given, and possible applications are noted.

621.318.423.011.3 52
Inductance of Air-Cored Single-Layer Cylindrical Coils—K. Schönbacher. (*Elektrotechnik* (Berlin), vol. 3, pp. 327–329; October, 1949.) A simple formula involving the ratio of the coil diameter d to its length l is derived from expressions given by Rayleigh and by Lorenz. For all practical cases, the use of correction formulas can be avoided by replacing l by $l+0.45d$.

621.318.572:621.396.645:537.311.33:621.315.59 53
A Transistor Trigger Circuit—H. J. Reich and R. L. Ungvary. (*Rev. Sci. Instr.*, vol. 20, pp. 586–588; August, 1949.) A trigger circuit may be formed by inserting a suitable resistor in the lead connected to the base of a transistor. Such circuits are in many ways superior to tube circuits; a circuit with a triggering time of 0.1 μs giving an output of 5–6 v was found to be stable at frequencies up to 1 Mc; stability at frequencies up to at least 10 Mc is probable. The circuit may be converted into a relaxation oscillator or a pulse generator.

621.392 54
A Network Bisection Theorem—V. D. Landon. (*RCA Rev.*, vol. 10, pp. 448–450; September, 1949.) It is shown how any given symmetrical ladder network may be split into two simpler networks such that the transfer factor of the whole is equal to half the product of the transfer factors of the parts, as here defined.

621.392:621.3.015.3 55
The Effect of Pole and Zero Locations on the Transient Response of Linear Dynamic Systems—J. H. Mulligan, Jr. (Proc. I.R.E., vol. 37, p. 1181; October, 1949.) Corrections to 2165 of 1949.

621.392.5:621.392.26† 56
Basis of the Application of Network Equations to Waveguide Problems—D. M. Kerns. (*Bur. Stand. Jour. Res.*, vol. 42, pp. 515–540; May, 1949.) A mathematical and fundamental paper. The solution of network problems can be effected by means of a matrix equation in which the constants are parameters characteristic of the components, and the variables are currents and voltages. The physical processes inside the components are not explicitly considered. A matrix equation is developed to enable this technique to be extended to waveguide and transmission-line problems; here tangential electric and magnetic fields replace the currents

and voltages as variables. Many of the results obtained are identical with those of the theory of quadripoles. In particular, a reciprocity theorem and Foster's reactance theorem are considered.

621.392.52 57
Error-Actuated Power Filters—G. Newstead and D. L. H. Gibbings. (PROC. I.R.E., vol. 37, pp. 1115–1119; October, 1949.) Voltages derived from the harmonic voltages of an ac source of impure waveform are applied through a negative-feedback path to cancel unwanted harmonics. The method could be extended to apply to a given band of frequencies. Large powers can thus be controlled by means of components of much lower rating. Design equations are given and the operation of a typical filter circuit is examined.

621.392.52 58
Smoothing Circuits: Part 2—Inductance-Capacitance—"Cathode Ray." (*Wireless World* vol. 55, pp. 418–422; November, 1949.) Continuation of 3371 of 1949. The calculation of hum voltages is discussed.

621.392.52 59
Mechanical Filters for Radio Frequencies—W. van B. Roberts and L. L. Burns, Jr. (*RCA Rev.*, vol. 10, pp. 348–365; September, 1949.) The theory of neck-type and slug-type filters is discussed. A simple type of band-pass filter composed of loosely coupled metal resonators with magnetostrictive drive and take-off has very sharp frequency discrimination and is readily constructed for frequencies up to $\frac{1}{2}$ Mc and bandwidth less than 3 per cent of the mid-frequency. The voltage gain of an amplifier stage using such a filter is generally considerably lower than that of a stage using electrical circuit coupling.

621.392.52:621.3.015.3 60
Transient Response of Filters—M. S. Corrington. (*RCA Rev.*, vol. 10, pp. 397–429; September, 1949.) The ideal low-pass filter is defined as one whose amplitude response is constant at all frequencies between zero and cut-off, and which falls off at a fixed rate beyond cut-off. The flat characteristic of such a filter is not necessarily desirable when the rise time and the amount of overshoot are important. The transient response of an ideal filter has been computed, and the results applied to the determination of the transient response for systems with selectivity curves composed of straight-line segments. Relations between the transient response of a low-pass filter and that of the corresponding high-pass filter are considered.

621.392.52.012.8 61
Generalized Theory of the Band-Pass Low-Pass Analogy—P. R. Aigrain, B. R. Teare, Jr., and E. M. Williams. (PROC. I.R.E., vol. 37, pp. 1152–1155; October, 1949.) Laplace transforms are used. Video-circuit equivalents are shown to exist for asymmetrical systems with low-level modulation as well as for symmetrical systems. Several illustrations are given.

621.392.6 62
Synthesis of Passive Networks with Any Number of Pairs of Terminals, Given Their Impedance or Admittance Matrices—M. Bayard. (*Bull. Soc. Franç. Élec.*, vol. 9, pp. 497–502; September, 1949.) The necessary and sufficient condition is established for a square symmetric matrix to be an impedance or an admittance matrix. The condition is that the quadratic form associated with the given matrix is a "positive real function" in the sense in which Brune uses the term. This result generalizes the theorem established by Brune for the impedance matrices of 2-pole networks (1932 Abstracts, p. 280) and by Gewertz for those of quadripoles (1934 Abstracts, p. 514). The greater part of the mathematical treatment included in the original paper presented at a meeting of the Société Française des Électriciens is here omitted.

621.396.611.1 63
Energy Fluctuations in a van der Pol Oscillator—N. Minorsky. (*Jour. Franklin Inst.*, vol. 248, pp. 205–223; September, 1949.) The behavior of the van der Pol oscillator, whose differential equation is

$$\ddot{x}-\epsilon(1-x^2)\dot{x}+x=0,$$

resembles that of the harmonic oscillator when ϵ is small; the very different behavior when ϵ is large is examined with the aid of data obtained by the method of isoclines.

The van der Pol oscillator has a prescribed energy content which does not depend on the initial conditions, whereas the harmonic oscillator can have any energy content. The van der Pol oscillator is nevertheless the best means available for physical realization of simple harmonic oscillation if ϵ is small. The perturbation method is used to determine the fluctuations of energy with time and azimuth for $\epsilon \ll 1$. For $\epsilon \gg 1$, the energy fluctuations of the van der Pol oscillator are quasi-discontinuous, and somewhat analogous to those of a pneumatic hammer.

621.396.611.1:621.3.015.33 64
Response of RC Circuits to Multiple Pulses—D. Levine. (PROC. I.R.E., vol. 37, pp. 1207–1208; October, 1949.) The voltage across the capacitor due to the application of several pulses in succession is determined by a generalized step-by-step process for any number of pulses. Only the well-known charge/discharge functions for a dc circuit are required. The voltage across the resistor can then be directly determined.

621.396.611.1.029.63 65
Tank Circuits as Resonators in Decimetre-Wave Technique—G. Megla. (*Elektrotechnik* (Berlin), vol. 2, pp. 305–312; November, 1948.) Detailed treatment of the properties of tank circuits, with illustrations of many different types and derivation of suitable formulas for the resonance frequency and resistance. Below a wavelength of about 10 cm the Q of tank circuits decreases, so that cavity resonators are preferable.

621.396.615:621-12 66
The Reciprocator—W. C. White and H. W. Lord. (*Electronics*, vol. 22, pp. 70–71; November, 1949.) A ring oscillator, comprising 2 one-shot multivibrators whose on and off periods can be varied independently between 0.2 and 1.5 seconds, energizes linear or rotary solenoids which generate reciprocating motion in which the distance/time relationship during the whole stroke can be controlled.

621.396.614.141.1/.2 67
On the Theory of Magnetron Barkhausen Oscillations—H. G. Möller. (*Elektrotechnik* (Berlin), vol. 3, pp. 129–133; May, 1949.) The mechanism of the production of Barkhausen oscillations in whole-anode magnetrons is similar to that for triodes. For magnetrons the period of the pendular electrons is shorter than that of the retraded electrons, the reverse being the case for triodes. In consequence of the different frequencies, phase focusing occurs. An explanation of the oscillation mechanism is put into a mathematical form and the approximate paths of the electrons, the dependence of the pendular frequency on the amplitude, the ac contribution to the space charge, and the excitation factor, are calculated. An upper limit for the efficiency can be found if it is assumed that the total pendular energy can be converted to oscillation energy.

621.396.615.17:621.317.755 68
Panoramic Sweep Circuits—C. B. Clark and F. J. Kamphoefner. (*Electronics*, vol. 22, pp. 111–114· November, 1949.) Brief details, with block diagrams, of 12 methods of obtaining sweep voltages for panoramic receivers, FM signal generators, and rf spectrum analyzers.

621.396.619.23:621.396.615.17 69
A Modulator Producing Pulses of 10^{-7} Second Duration at a 1-Mc/s Recurrence Frequency—M. G. Morgan. (PROC. I.R.E., vol. 37, p. 1178; October, 1949.) Correction to 2185 of 1949.

621.396.645 70
Musician's Amplifier—D. Sarser and M. C. Sprinkle. (*Audio Eng.*, vol. 33, pp. 11–13, 55; November, 1949.) An adaptation of the Williamson circuit (2715 of 1945 and 3101 of 1949). Williamson's output-transformer performance specifications are satisfactorily met by the Peerless transformer Type S-265Q. Construction and performance details are discussed and shown graphically. Quality is outstanding.

621.396.645 71
Amplification by Direct Electronic Interaction in Valves without Circuits—P. Guénard, R. Berterottiére, and O. Doehler. (*Bull. Soc. Franç. Élec.*, vol. 9, pp. 543–549; October, 1949.) *See* 2977 of 1949.

621.396.645 72
Simplified Preamplifier Design—H. T. Sterling. (*Audio Eng.*, vol. 33, pp. 16–17, 45; November, 1949.) Unnecessary complications may be avoided by making practical compromises in performance requirements. A modified Pickering circuit is recommended. See also 1608 of 1948 (Burwen).

621.396.645:539.17 73
Fast Pulse-Amplifiers for Nuclear Research—W. C. Elmore. (*Nucleonics*, vol. 5, pp. 48–55; September, 1949.) Discussion of the limitations affecting the speed of response of amplifiers without feedback, and of the design of amplifiers whose speed of response approaches the theoretical limit. In each stage of the amplifier, the output signal is assumed to rise monotonically to its final steady value for a step-voltage input. The relations between the over-all rise time of an amplifier and the rise times of individual stages are considered. Characteristics of 11 tubes useful for pulse amplifiers are tabulated and discussed.

621.396.645:578.088.7 74
The Design and Construction of an Amplifier for Bio-Electric Recording—R. T. Jamieson. (*Trans. S. Afr. Inst. Elec. Eng.*, vol. 40, pp. 204–212; September, 1949. Discussion, p. 212.) Requirements of such amplifiers are discussed. An ac mains-operated amplifier is described. The pre-amplifier has a gain of 74 db and consists of 2 coupled cathode stages followed by a cathode follower with gas-tube coupling to the input of the high-level amplifier, which is a 3-stage resistance-coupled amplifier with cathode-follower output and a gain of 66 db. Full-scale deflection on the cro is obtained for a 10-μV input signal. The two amplifiers should be isolated in places reasonably free from electrical noise.

621.396.645:621.385 75
Operation of Output Valves in High-Power Public-Address Amplifiers—N. L. Bezladnov. (*Radiotekhnika* (Moscow), vol. 4, pp. 8–17; July and August, 1949. In Russian.) Continuation of 2662 of 1949. The efficiency of the final stage of such amplifiers can be improved by (a) anode loading, (b) use of a limiting device. In case (a), the average anode dissipation under dynamic operating conditions, not the maximum dissipation, should be regarded as determining maximum permissible anode loading. In case (b), the transmission power can be increased without increasing nonlinear distortion.

The theory of both methods is discussed and experimental curves are shown.

621.396.645.37:518.4 76
Calculator and Chart for Feedback Problems—J. H. Felker. (PROC. I.R.E., vol. 37, pp. 1204–1206; October, 1949.)

621.396.645.37.001.4:621.3.016.352:
621.3.015.3 77

Examination of Amplifier Stability by Applying a Sudden D.C. Test Voltage—B. Carniol. (*Tesla Tech. Rep.* (Prague), pp. 11–20; March, 1949.) Nyquist's theorem and the corresponding characteristic curve for feedback amplifiers are discussed and practical methods for determining the characteristic for any particular amplifier are described in detail. The computation accuracy depends on the exactness with which the phase characteristic can be represented by a series of straight lines. The advantages of the method of testing amplifier stability by studying the response to a transient voltage or square wave are considered. The effect of circuit variations in enhancing or reducing transient oscillations in the output of various amplifiers is illustrated by a series of oscillograms. A Nyquist characteristic for one of the amplifiers shows the existence of a critical frequency near which a definite tendency to instability exists. See also 218 below.

621.396.645.371:621.3.015.3 78

Transient Phenomena in Wide-Band Feedback Amplifiers—O. B. Lur'e. (*Zh. Tekh. Fiz.*, vol. 19, pp. 952–972; August, 1949. In Russian.) Negative-feedback amplifiers should be widely used in video circuits although they give somewhat lower gain than that obtained with the usual distortion-correcting circuits. Formulas are derived for determining circuit parameters when given (a) the time for reaching the steady state, (b) the maximum permissible overshoot.

621.396.662:621.396.61 79

A New Type of V.H.F. Tank Design—B. E. Parker. (*FM-TV*, vol. 9, pp. 14–15; October, 1949.) A balanced 2-wire section of transmission wire is tuned by varying its distance from a flat conductor. This affects the series inductance and the shunt capacitance, and hence the surge impedance. This arrangement has advantages in a push-pull output circuit, including the fact that the tuning element is at earth potential.

621.396.662:621.396.61:513.76 80

Bilinear Transformations Applied to the Tuning of the Output Network of a Transmitter—K. S. Kunz. (Proc. I.R.E., vol. 37, pp. 1211–1217; October, 1949.) The theory of bilinear transformations in a complex plane, by which circles are transformed into circles, is used to determine the actual value C of the secondary capacitance for (a) resonance of the secondary circuit, (b) a maximum or minimum of the average anode current, (c) maximum coupled-circuit efficiency. (b) and (c) require values of C differing widely from that required for (a), except for the case of a series load in the secondary circuit. Although the values of C required for (b) and (c) may differ considerably, the difference is small if the Q of the primary coil is large and the coupling coefficient very small.

621.396.662.2 81

A New Decade Inductor—H. W. Lamson. (*Gen. Radio Exp.*, vol. 24, pp. 1–8; July, 1949.) For the range 1 mh–10 h intended for use at audio and the lower ultrasonic frequencies. Each decade unit consists of four inductors whose inductances are in the ratio 1:2:2:5, with an over-all es shield and a special switch for connecting the inductors in series as required. Multi-layer uniformly wound toroidal coils are used. These have a banked winding to minimize distributed capacitance and can be stacked in a compact unit with negligible mutual inductance. The stabilized Mo-permalloy dust cores have low eddy-current and hysteresis losses. Construction details and electrical properties are discussed.

GENERAL PHYSICS

530.12:531.18 82

The Special Relativity Theory—E. Kübler. (*Elektrotechnik* (Berlin), vol. 2, pp. 323–327; November, 1948.) A simple explanation, by an electrician for electricians, of the electrodynamic consequences of this theory.

534.26+535.42 83

On the Theory of Diffraction by an Aperture in an Infinite Plane Screen: Part 1—H. Levine and J. Schwinger. (*Phys. Rev.*, vol. 74, pp. 958–974; October 15, 1948.) The case of a scalar plane wave is considered. The wave function at an arbitrary point in space is expressed in terms of its values in the aperture, and constructed so as to vanish on the screen, in accordance with the assumed boundary condition. An integral equation is obtained, by means of which the amplitude of the diffracted spherical wave at large distances from the aperture can be expressed in a form which is stationary with respect to small variations of the aperture fields arising from a pair of incident waves. This expression is independent of the scale of the aperture fields. The transmission cross section of the aperture for a plane wave is simply related to the diffracted amplitude observed in the direction of incidence. The method is applied to the case of a wave incident normally on a circular aperture, for which exact results are available by using suitable trial aperture fields in the variational method, approximate but sufficiently accurate results are obtained for the diffracted amplitude and transmission cross section over a wide range of frequencies. See also 3294 of 1948 (Spence).

534.26+535.42 84

On the Transmission Coefficient of a Circular Aperture—C. J. Bouwkamp, H. Levine, and J. Schwinger. (*Phys. Rev.*, vol. 75, pp. 1608–1609; May 15, 1949.) Comment on 83 above and the authors' reply.

535.42:538.566 85

New Solution of the Problem of the Diffraction of Electromagnetic Waves by a Plane Perfectly Conducting Screen—J. P. Vasseur. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 586–587; September 19, 1949.) Using the same notation as in a previous note (3124 of 1949), a rigorous expression of Babinet's principle is derived by the help of integral equations, the integration extending over the surface of the screen. This expression of Babinet's principle has been given by H. G. Booker in an unpublished paper. See also 1335 of 1947 (Booker).

537.311.62 86

Skin Effect—C. Vazaca. (*Radio Tech. Dig.* (Franç), vol. 3, pp. 175–184, 223–233, and 307–316; June–October, 1949.) Reprint of article in *Bull. Nat. Recherches Tech.* (Roumanie), vol. 3, 1948. The concentration of the current in the surface layers of a conductor carrying ac is quite analogous to the concentration of magnetic field in the surface layers under the action of a longitudinal alternating magnetic field, owing to eddy-current effects. Starting from Maxwell's equations, formulas are developed for the case of a long cylindrical conductor, giving the law of distribution within the conductor of the current density and total current, and also of the magnetic field and total magnetic flux. Formulas are also derived for the impedance and the complex reluctance, the effective depth of penetration of the current and of the magnetic field, and the energy dissipated per unit length or volume of the conductor.

537.521.7 87

The Positive Streamer Mechanism of Spark Breakdown—W. Hopwood. (*Proc. Phys. Soc.*, vol. 62, pp. 657–664; October 1, 1949.)

537.525:538.551.25 88

Theory of Plasma Oscillations. A. Origin of Medium-Like Behavior—D. Bohm and E. P. Gross. (*Phys. Rev.*, vol. 75, pp. 1851–1864; June 15, 1949.) A theory of electron oscillations of an unbounded plasma of uniform ion density is developed, taking into account the effects of random thermal motions, but neglecting collisions. The frequencies are determined at which a plasma can undergo organized steady-state small oscillations to which a linear approximation applies. Organized oscillations of wavelength smaller than the Debye length for the electron gas are not possible. For a given wavelength, a plasma can oscillate with arbitrary frequency, but the frequencies not satisfying a "steady-state dispersion relation" describe motions in which, after some time, there is no contribution to large-scale averages. The treatment is extended to the case of large steady-state oscillations for which the equations are nonlinear, and the general nature of solutions for this case is discussed. See also 89 below.

537.525:538.551.25:621.396.822 89

Theory of Plasma Oscillations. B. Excitation and Damping of Oscillations—D. Bohm and E. P. Gross. (*Phys. Rev.*, vol. 75, pp. 1864–1876; June 15, 1949.) The theory discussed in 88 above is extended to include the effects of collisions and of special groups of particles having well-defined ranges of velocities. A wave tends to be damped by collisions in a time of the order of the mean time between collisions. Groups of particles having speeds far above the mean thermal speed may cause instability only limited by effects neglected in the linear approximation. In the absence of plasma oscillations, any beam of well-defined velocity is scattered by the individual plasma electrons acting at random, but the scattering is much greater when all the particles oscillate in unison. The beams are thus scattered by the oscillations they produce. This type of instability may be responsible for solar and cosmic rf noise.

538.12 90

Production of Intense Magnetic Fields by means of Pulses.—G. Raoult. (*Ann. Phys.* (Paris), vol. 4, pp. 369–421; July and August, 1949.) By means of recurrent voltage pulses, a suitable commutator and a grid-controlled Hg rectifier, a capacitor is charged and then discharged through a coil about 100 times per second. In this way magnetic field pulses of roughly sinusoidal shape and of duration about 15μs have been produced with a peak power of the order of 90 Mw and a mean power of 850 w. The maximum field strength was 52,000 gauss in a useful volume of about 3 cm^3. Applications to phenomena of polarization rotation and magnetic double refraction in CS_2 are described.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523:621.396.9 91

Astronomy—J. S. Hey. (*Mon. Not. R. Astr. Soc.*, vol. 109, pp. 179–211; 1949. Bibliography, pp. 211–214.) General survey of (a) radio detection of meteors and reflections from the moon, (b) solar radiation, and (c) galactic radiation.

523.746.5 92

The Probable Behavior of the Next Sunspot Cycle—W. Gleissberg. (*Astrophys. Jour.*, vol. 110, pp. 90–92; July, 1949.) The probability laws of sunspot variations, which have yielded successful predictions for the present sunspot cycle, suggest that a steep ascent to a very high maximum is highly probable in the next cycle, and that the period of low activity preceding the next cycle will be extremely short.

550.385"1949.01.24/.26" 93

Recurrence Features of the Magnetic Storm of 1949, January 24–26—(*Observatory*, vol. 69, pp. 195–196; October, 1949.) Great magnetic storms, beginning with a "sudden commencement" and associated with solar flares, do not usually show any recurrence connected with the solar rotation. This particular storm, however, did recur with sudden commencements on February 21, 1949 and March 21, 1949. No positive evidence of flares associated with the last two storms is available. These storms were of the polar type, and were associated with auroras. Magnetograms are reproduced and compared.

551.51 94
The Earth's Upper Atmosphere—D. R. Bates. (*Mon. Not. R. Astr. Soc.*, vol. 109, pp. 215–242; 1949. Bibliography, pp. 242–245.) Known facts about the constitution, density and temperature at various levels are summarized, and methods of exploring the upper atmosphere are surveyed. The ionized layers and auroras are also discussed.

551.510.535 95
Ionization in the Earth's Upper Atmosphere—(*Observatory*, vol. 69, pp. 185–191; October, 1949.) Report of a discussion held at Manchester University on July 2, 1949.

551.557.7:621.396.9 96
A Radio Method of Measuring Winds in the Ionosphere—S. N. Mitra. (*Proc. IEE* (London), vol. 96, pp. 441–446; September, 1949.) A pulse transmitter operating at a frequency of about 4 Mc is used in conjunction with 3 receivers at the corners of a right-angled triangle with sides of the order of 100 m. A single magneto-ionic component received at one point will show fading only if the ionosphere changes either by random motion of irregularities or by steady motion of the ionosphere as a whole. A uniform drift of the ionosphere would cause the diffraction pattern to move past a receiver with twice the velocity of the drift. Wind velocities can thus be deduced when the records at the 3 receivers are similar but displaced in time. Results thus obtained are compared with those of previous measurements by other methods.

551.594.6 97
Large-Scale Variation of the Level of Atmospherics during the Antarctic Cruise of the Commandant Charcot—R. Bureau and M. Barré. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 626–627; September 26, 1949.) The recorder used on board the Commandant Charcot has been calibrated by Carbenay's method (2902 of 1948). A map is given showing, for the various sections of the whole course, the operational threshold in maxwells per meter. Longitude and latitude effects are clearly indicated and the difference between the threshold sensitivities for January and March in Australian waters indicates also a seasonal effect. See also 3429 of 1949.

551.594 98
Atmospheric Electricity [Book Review]—J. A. Chalmers. Publishers: Oxford University Press, Oxford, 1949, 175 pp., 15s. (*Proc. Phys. Soc.*, vol. 62, pp. 665–666; October 1, 1949.)

LOCATION AND AIDS TO NAVIGATION

621.396.9 99
Radar—(*Bull. Soc. Franç. Élec.*, vol. 9, p. 532; October, 1949.) In Quillet's encyclopedia the following entries may be found:
Radar.—En Perse: Membre d'une milice créée pour protéger les voyageurs. [In Persia: Member of a militia established for the protection of travelers.]
Radarie —Droit persan payé par les marchandises sous la protection des radars. [Persian toll paid on merchandise under the protection of radars.]

621.396.9:061.4 100
Radar Research and Development Establishment [Malvern]—(*Engineer* (London), vol. 188, p. 373; September 30, 1949.) A brief survey of work in progress at this establishment, based on a visit made on a recent open day. See also *Nature* (London), vol. 164, pp. 740–741; October 29, 1949.

621.396.9:621.385.832:535.371.07 101
Radar Screens—de la Pinsonie. (*See* 267.)

621.396.933:526.92 102
Distance-Measuring Equipment for Aircraft Navigation—V. D. Burgmann. (*Proc. IEE* (London), vol. 96, pp. 395–402; September, 1949.) An airborne light-weight radar set works in conjunction with a responder beacon on the ground. The radar transmitter sends out a stream of pulses which are returned by the beacon at a slightly different frequency. The time interval between the outgoing and corresponding return pulse is measured automatically and shown on a meter calibrated in miles. The design of the airborne set and results obtained in trials on an air route are discussed. Well-established techniques are used.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.37 103
Review of the Interpretations of Luminescence Phenomena—F. E. Williams. (*Jour. Opt. Soc. Amer.*, vol. 39, pp. 648–654; August, 1949.)

535.37 104
A Survey of Present Methods Used to Determine the Optical Properties of Phosphors—W. B. Nottingham. (*Jour. Opt. Soc. Amer.*, vol. 39, pp. 641–647; August, 1949.)

535.37 105
The Intensity Dependence of the Efficiency of Fluorescence of Willemite Phosphors—W. Hoogenstraaten and F. A. Kröger. (*Physica's Grav.*, vol. 15, pp. 541–556; July, 1949. In English.)

535.37 106
Temperature Quenching and Decay of Fluorescence in Zinc and Zinc-Beryllium Silicates Activated with Manganese—F. A. Kröger and W. Hoogenstraaten. (*Physica's Grav.*, vol. 15, pp. 557–568; July, 1949. In English.) Continuation of 3156 of 1949 and 105 above.

535.37:546.412.84 107
Optical Properties of Calcium Silicate Phosphors—F. J. Studer and G. R. Fonda. (*Jour. Opt. Soc. Amer.*, vol. 39, pp. 655–660; August, 1949.)

535.37:546.471.61 108
A Correlation between Cathodoluminescence Efficiency and Decay [of Mn-activated ZnF_2] as a Function of Temperature—R. H. Bube. (*Jour. Opt. Soc. Amer.*, vol. 39, pp. 681–684; August, 1949.)

535.37:546.472.21 109
Sodium and Lithium as Activators of Fluorescence in Zinc Sulfide—F. A. Kröger. (*Jour. Opt. Soc. Amer.*, vol. 39, pp. 670–672; August, 1949.)

535.37:546.472.21 110
The Luminescence of Zinc Sulfide Activated by Lead—N. W. Smit and F. A. Kröger. (*Jour. Opt. Soc. Amer.*, vol. 39, pp. 661–663; August, 1949.)

537.311.31+536.212.2]:546.78 111
Thermal and Electrical Resistance of a Tungsten Single Crystal at Low Temperatures and in High Magnetic Fields—J. de Nobel. (*Physica,'s Grav.*, vol. 15, pp. 532–540; July, 1949. In English.)

537.312.62:621.396.622 112
Detection at Radio Frequencies by Superconductivity—Lebacqz, Clark, Williams, and Andrews (*See* 223.)

538.221 113
Magnetic Materials for Electrical Power Plant—F. Brailsford. (*Engineering* (London), vol. 168, pp. 293–296; September 16, 1949.) Long summary of paper read before the British Association. The progress in steel and sheet-making processes is reviewed. Different steels are compared; the need for improving magnetic quality is stressed. Methods outlined include directional cold-rolling technique and a laboratory method for preferred orientation in silicon-iron crystals. For other summaries see 3442 of 1949 and *Electrician*, vol. 143, pp. 737–738; September 2, 1949. For discussion, see *Electrician*, vol. 143, pp. 893–894; September 16, 1949.

549.514.51 114
The Quartz Crystal as Raw Material for High-Frequency Technics—H. Neels. (*Elektrotechnik* (Berlin), vol. 3, pp. 318–322; October, 1949.) Discussion of the various forms of natural quartz, methods of selection by visual inspection, and optical and electrical tests for twinning.

549.514.51 115
A New Crystal Cut for Quartz with Zero Temperature Coefficient—E. J. Post. (*Tijdschr. ned. Radiogenoot.*, vol. 14, pp. 147–157; September, 1949.) A rotated y-cut making an angle of about $+36°$ with the z-axis. It is suitable for the frequency range 400–1000 kc and the term HT-cut is proposed. Resonators can be mounted by soldering springs at the nodal points. Spurious responses are absent in the range within 10 per cent above or below the main resonance frequency.

549.514.51:621.392.52 116
Crystals for Electrical Filters—R. Taylor, R. Bechmann, and A. C. Lynch. (*Research* (London), vol. 2, pp. 414–417; September, 1949.) The equivalent circuit of a quartz crystal is discussed, and factors that must govern the choice of suitable crystals for practical applications are considered. A crystal may have as many as 18 piezoelectric constants, 6 permittivities, and 21 elastic constants, all of which may have different temperature coefficients. The highest piezoelectric constants for various classes of piezoelectric crystal are tabulated. Temperature coefficients, ferroelectric properties, methods of growing crystals, etc., are briefly discussed. The ideal piezoelectric material has yet to be found.

621.315.59+621.314.6 117
Semiconductor Rectifiers—(*See* 240.)

621.315.59+621.314.63 118
Semi-Conductors and Rectifiers—N. F. Mott. (*Proc. IEE* (London), vol. 96, pp. 253–260; September, 1949.) Full paper: summary noted in 2814 of 1949.

621.315.59+621.315.61 119
New Dielectric Materials: Semi-Conductors, Plastics and Silicones—(*Electrician*, vol. 143, pp. 735–736; September 2, 1949. Discussion, *ibid.*, vol. 143, pp. 893–894; September 16, 1949.) Summary of paper presented before the British Association by R. W. Sillars, entitled "Insulating and semi-conducting materials." Another summary noted in 3442 of 1949. See also 3168 of 1949 (Jones, Scott, and Sillars).

621.315.612 120
Titanates with High Dielectric Constant—H. Sachse. (*Z. Angew. Phys.*, vol. 1, pp. 473–484; August, 1949.) A comprehensive review of the results of measurements, made in various countries, of the properties of a wide range of titanate materials. 57 references are given.

621.315.612:621.317.37.029.6 121
The Measurement of the Dielectric Properties of High-Permittivity Materials at Centimetre Wavelengths—J. G. Powles and W. Jackson. (*Proc. IEE* (London), vol. 96, pp. 383–389; September, 1949.) Results of room-temperature measurements on ceramic specimens of the titanates of Mg, Ca, Sr, and Ba at 1.5 Mc and at 9,450 Mc, and on $BaTiO_3$ at 24,000 Mc are discussed. $BaTiO_3$ is unique among these materials, because a considerable fall in permittivity and a large increase in tan δ occur at the higher frequencies. Variable-temperature measurements on $BaTiO_3$ show that the crystallographic change associated with the Curie point at 120°C affects the permittivity at 9,450 Mc and at 1.5 Mc in an

analogous manner. Measurements at the same frequencies on (Ba, Sr) titanate compositions are also described; the composition 56 per cent $BaTiO_3$:44 per cent $SrTiO_3$ has a permittivity of 760 and tan $\delta = 0.02$ at 9,450 Mc and 20°C. The three methods of measuring high permittivities at cm developed in the course of this work are described.

621.315.614:621.317.372 122
Electrical Testing of Capacitor Paper—Endicott. (*See* 144.)

621.315.614.62/.63:621.319.4 123
Metallizing Paper for Capacitors—H. G. Wehe. (*Bell Lab. Rec.*, vol. 27, pp. 317–321; September, 1949.) A German wartime metal-vapor spraying process, developed by the Bell Telephone Laboratories. For a given capacitance the metallized-paper capacitor is much smaller than the corresponding metal-foil capacitor, and has the advantage of self-healing of punctures through the insulation.

621.315.615.011.5:537.226.2 124
A Contribution to the Theory of the Dielectric Constant of Polar Liquids—T. G. Scholte. (*Physica,'s Grav.*, vol. 15, pp. 437–449; July, 1949. In English.)

621.315.615.2.011.5:537.226.2 125
The Absolute Dielectric Constant of Benzene—F. van der Maesen. (*Physica,'s Grav.* vol. 15, pp. 481–483; July, 1949.) The dielectric constant is appreciably reduced by prolonged drying over Na. This confirms the conclusion of Hartshorn and Oliver (*Proc. Roy. Soc. A.*, vol. 123, pp. 664–685; 1929) that moisture is the only impurity of importance in dielectric-constant measurements. Values obtained differ by less than 0.1 per cent from the generally-accepted standard values. The method of purification is discussed.

621.315:616 126
Ethoxylines—E. Preiswerk and C. Meyerhans. (*Elec. Mfg.*, vol. 44, pp. 78–81, 166; July, 1949.) Properties and applications are discussed. The first resins of this type were supplied under the trade name Araldite; for another account see 3450 of 1949 (Moss).

621.315.616:546.287 127
Recent Progress Made in Silicone Rubber Materials—C. E. Arntzen and R. D. Rowley. (*Materials and Methods*, vol. 30, pp. 73–76; October, 1949.) The chemical nature of silicones, mechanical and physical properties of various silastic rubbers, and applications are discussed. Silicone rubbers are much less affected by temperatures of the order of 300°F than natural rubbers.

621.315.618.011.5:537.226.2 128
The Absolute Dielectric Constant of Gases at Pressures of 0–80 Atm. at 25°C—J. Clay and F. van der Maesen. (*Physica,'s Grav.*, vol. 15, pp. 467–480; July, 1949. In English.) Experimental values for air, He, Ar, and CO_2 are tabulated and discussed. A heterodyne beat method is used for determining the absolute values of small differences of capacitance by means of Clay's cylindrical measuring capacitor (141 below).

621.318.22:538.652 129
The Magnetostriction of Anisotropic Permanent Magnet Alloys—M. McCaig. (*Proc. Phys. Soc.*, vol. 62, pp. 652–656; October 1, 1949.) Blocks of permanent magnet alloys of the system FeNiAlCoCu were prepared with columnar crystals. The magnetostriction was measured in various directions before and after heat treatment in a magnetic field. The results agree well with those predicted theoretically, and so confirm the postulated crystal structure.

621.775.7 130
Sintering of Iron Powders—H. Bernstorff. (*Metal Treat.*, vol. 15, pp. 85–89; Summer, 1948.) Tabulated results of experimental studies in the Degussa laboratories on the sintering of DPG Schleuder iron powder are used to derive curves showing tensile strength and elongation as a function of sintering time. For the higher sintering temperatures the curves approach constant values. At lower temperatures distinct minima are found for both strength and elongation.

666.1.037.5 131
Glass-to-Metal Seals—(*Metal Ind.* (London), vol. 75, pp. 263–266 and 292–293; September 30 and October 7, 1949.) The relative merits of various metals and glasses used for sealing and various types of seal are discussed characteristics of certain glasses are tabulated. For many purposes, BTH C40 borosilicate glass sealed to Nilo-K, an alloy consisting of 54 per cent Fe, 31 per cent Ni and 15 per cent Co, is the best combination now available. No great skill is required in preparing or manufacturing the seal; pre-beading is unnecessary. Seals with diameters up to at least 5 inches have been successfully made with these materials. The nature of stresses in seals is briefly discussed.

MATHEMATICS

512.974:621.3 132
Vectorial Space. Region of Representation for Electrotechnics—G. Nasse. (*Bull. Soc. Franç. Élec.*, vol. 9, pp. 459–474; September, 1949.) The concept of the region of representation, limited in classical teaching to the Fresnel plane, is shown to be capable of a natural extension leading to the vectorial space of geometers, a metric space capable of representing under one and the same structure the properties of monophase and polyphase systems, and even those of their transitory states.

681.142 133
The University of Manchester Universal High-Speed Digital Computing Machine—T. Kilburn. (*Nature* (London), vol. 164, pp. 684–687; October 22, 1949.) The binary system is used, and numbers up to 2^{40} can be accommodated. Operations which the machine can perform and "addresses" where numbers can be stored also have numbers. "Instructions" take the form of numbers which specify that a certain operation is to be performed on the contents of a certain address. An electronic storage unit with a capacity of 5,120 digits is included; this type of storage was discussed in 2258 of 1949 (Williams and Kilburn). A magnetic storage unit is also available in which the numbers stored are less accessible but which has a capacity of 40,960 digits a similar storage system was described by Booth (3179 of 1949). Special circuits are provided to perform addition, subtraction, multiplication, and certain logical operations, but a "sub-program" is used for other operations, including division, which occur less frequently. The component elements of the computing circuits used are similar to those used in video-pulse circuits in radar (see 1874 of 1948). One type of adding circuit is considered in detail. The "control" is an electronic storage tube, holding two numbers; its operation is described. Arrangements are made to enable the machine to perform or omit an instruction according to the nature of the result obtained at a particular stage of the calculation.

681.142:621.385.032.212 134
Polycathode Glow Tube for Counters and Calculators—Lamb and Brustman. (*See* 275.)

MEASUREMENTS AND TEST GEAR

531.764.5 135
The Development of the Quartz Clock—W. A. Marrison. (*Elektrotechnik* (Berlin), vol. 3, pp. 311–316; October, 1949.) Condensed version of the paper noted in 762 of 1949.

621.317.2 136
New High-Voltage Engineering Laboratory—J. H. Hagenguth. (*Gen. Elec. Rev.*, vol. 52, pp. 9–14; September, 1949.) The laboratory was built for development and research work on apparatus used for power transmission and distribution. Test power for ac hv tests is supplied by one 1,000-kva generator and two 500-kva generators. Two 500-kw 500-v dc generators are also available. For studying transient voltages, such as those due to lightning, under controlled conditions, two 5,100-kv impulse generators are provided. These can produce a peak voltage of 7,500 kv to ground at a current of 33 ka when operated individually, or 66 ka when connected in parallel. Five 350-kv 1,000-kva test transformers can be connected in cascade to deliver 1,750 kv (rms) to ground. A surge current generator is also available, capable of producing 260 ka at 50, 100, or 150 kv.

621.317.3.011.5:621.365.55: 137
Dielectric Loss with Changing Temperature—J. B. Whitehead and W. Rueggeberg. (*Elec. Eng.*, vol. 68, p. 874; October, 1949.) Summary only. A method of measuring the properties of a dielectric while its temperature is changing has been developed and applied to certain thermosetting materials subjected to hf heating. The presence of absorbed moisture in small amounts has a pronounced effect on the dielectric properties of such materials. The two important phenomena are dielectric absorption and some form of molecular polarization; their relative importance depends on the type of heating cycle.

621.317.3.029.3 138
Audio-Frequency Measurements—W. L. Black and H. H. Scott. (PROC. I.R.E., vol. 37, pp. 1108–1115; October, 1949.) Discussion of the theory involved in making measurements of gain, frequency response, distortion, and noise, with particular reference to high-gain systems. Techniques of measurement and factors affecting accuracy are also discussed. See also 147 and 148 below.

621.317.32 139
Problems Related to Measuring the Field Strength of High-Frequency Electromagnetic Fields—R. Truell. (PROC. I.R.E., vol. 37, pp. 1144–1147; October, 1949.) Continuation of 429 of 1949, with more detailed treatment of the parallel-plate field and extension to the field in a rectangular cavity. In both cases, the field amplitude can be expressed quite simply in terms of easily measurable parameters, if certain velocity conditions are satisfied by the electron beam, and the cyclotron frequency $|e|H/mc$ is equal to the frequency of the em field.

621.317.32:551.594 140
A New Method of Measuring the Vertical Electric Field in the Atmosphere—J. A. Chalmers. (*Jour. Sci. Instr.*, vol. 26, pp. 300–301; September, 1949.) A metal ball rolls down an inclined earthed gutter and drops into an insulated collector, which thus acquires a potential proportional to the atmospheric electric field. This potential is amplified and measured by a galvanometer.

621.317.335.2†:621.319.4 141
Accurate Determination of the Absolute Capacity of Condensers: Part 2—J. Clay. (*Physica,'s Grav.*, vol. 15, pp. 484–488; July, 1949. In English.) Results are given of new precision measurements of the dimensions of three variable steel capacitors from which their capacitances are deduced. Values obtained agree to within 1 part in 3,000 with those found by es and oscillation methods, and the absolute value is believed to be accurate within 0.01 per cent. For part 1 see 4175 of 1936. These capacitors were used for the measurements of dielectric constant discussed in 125 and 128 above.

621.315.612:621.317.37.029.6 142
The Measurement of the Dielectric Proper-

ties of High-Permittivity Materials at Centimetre Wavelengths—Powles and Jackson. (*See* 121.)

621.317.372+621.317.334 143
Q and L Measurements—W. T. Cocking. (*Wireless World*, vol. 55, pp. 449–453; November, 1949.) When a Q meter is not available, the frequency or capacitance detuning methods here described form convenient ways of determining Q and also the inductance and self-capacitance of coils; Q is obtained by calculations involving measured frequencies or capacitances. Factors limiting accuracy are discussed.

621.317.372:621.315.614 144
Electrical Testing of Capacitor Paper—H. S. Endicott. (*Gen. Elec. Rev.*, vol. 52, pp. 28–35; September, 1949.) Equipment is described for measuring the dielectric constant and power factor of dry, unimpregnated capacitor paper. The sample is not removed from the drying chamber. The relation between the power-factor/temperature curves of dry and impregnated paper is discussed. The time necessary for a test has been reduced from 3 weeks to 2 days.

621.317.7+621.38+621.396.69]:061.4 145
Radiolympia Review—(*Wireless World*, vol. 55, pp. 428–444; November, 1949.) An illustrated review of various exhibits, namely (a) television equipment, (b) broadcast receivers, (c) sound reproducing equipment, (d) components and accessories, (e) communication equipment, (f) testing and measuring gear, (g) scientific, industrial, and medical apparatus. See also 3472 of 1949.

621.317.7.001.4 146
Operation and Care of Circular-Scale Instruments: Parts 1 and 2—J. Spencer. (*Instruments*, vol. 21, pp. 621–632 and 715–724; July and August, 1948.) A fully illustrated discussion of the mechanism of such instruments, and dismantling and reassembling procedure for repair, showing the principles of operation of Westinghouse and General Electric moving-coil dc meters Types KX24 and DB-12, and moving-iron ac meters Types KA24 and AB-12. Calibration adjustments, maintenance, and repair of various defects, from simple frictional defect to component failure, are also considered.

621.317.7.029.3 147
Audio Frequency Measurements: Part 1—W. L. Black and H. H. Scott. (*Audio Eng.*, vol. 33, pp. 13–16, 43; October, 1949.) Paper presented at a joint IRE-RMA meeting at Syracuse, N. Y. Definitions and minimum standards for the audio facilities of a broadcasting system are specified in R.M.A. Standard TR-105A, published in May, 1948. The technical background which is the basis for this specification is here surveyed and measurement methods are discussed thoroughly. A test circuit for gain measurements is described. The use of calibrated adjustable attenuators is recommended. Sources of error include mismatch and variation in the degree of earthing. The measurement of bridging gain is considered. Part 2: 148 below.

621.317.7.029.3 148
Audio Frequency Measurements: Part 2—W. L. Black and H. H. Scott. (*Audio Eng.*, vol. 33, pp. 18–19, 50; November, 1949.) Continuation of 147 above. Determination of harmonic distortion at the output, and intermodulation and noise measurements are discussed. See also 138 above.

621.317.7.088.22 149
The Ratio of Electrical to Friction Torques in Indicating Instruments—G. Szabó. (*Jour. Sci. Instr.*, vol. 26, pp. 301–304; September, 1949.) Factors affecting this ratio at its most favorable value are discussed. The shock-acceleration and shock-velocity are computed for the range of deformations where pivots and jewels are not damaged.

621.317.71:621.385.5 150
The Use of Multigrid Tubes as Electrometers—J. R. Prescott. (*Rev. Sci. Instr.*, vol. 20, pp. 553–557; August, 1949.)

621.317.715 151
The Galvanometer as a Built-In Component—G. Spiegel. (*Elec. Mfg.*, vol. 42, pp. 80–83; July, 1948.) Both the jewelled and the taut-suspension types of galvanometer are considered. The latter have much greater sensitivities and longer periods.

621.317.72 152
Compensator—A Device for Measurement of Alternating Currents within a Broad Frequency Range—V. Hlavsa. (*Tesla Tech. Rep.* (Prague), pp. 4–10; March, 1949.) For measurements at frequencies between 20 cps and 20 kc. One of two synchronous generators supplies current to the object under test and the voltage developed across it is balanced by the output of the other generator. The magnitude and phase of this output voltage are controlled by means of a potentiometer and a phase shifter, both of which are calibrated. Design details and applications of the equipment are discussed.

621.317.726:621.396.615.17 153
Measurement of the Time Constants of Peak Voltmeters Intended for the Study of Transient Phenomena—F. dr Clerck. (*HF* (Brussels), no. 3, pp. 81–87; 1949.) For measuring the time constant of the indicating instrument a rectangular pulse is used of duration variable and of the order of the time to be measured. For the time of charge of the electrical circuit a sine-wave pulse, symmetrical with respect to zero and with a rectangular envelope, is used, the pulse duration is again variable. The discharge time requires the sudden interruption, for a controlled time, of a sine wave. A generator for producing these three types of pulse is described and also a method of using it to enable the times to be measured on the screen of a cro. Other applications of the equipment are mentioned.

621.317.729:532.517.2 154
Fields from Fluid Flow Mappers—A. D. Moore. (*Jour. Appl. Phys.*, vol. 20, pp. 790–804; August, 1949.) Fluid flow within the streamline range, made visible by means of potassium permanganate crystals, can be used to simulate es, em, and other fields. Flow takes place between a flat slab of plaster-of-Paris, provided with suitable sources and sinks, and a parallel sheet of plate glass. Setting-up techniques for particular problems are described. Two-dimensional fields can in general be set up easily; some symmetrical three-dimensional fields can also be simulated. Photographs of results obtained are included and discussed.

621.317.73 155
Impedance Measurements with Directional Couplers and Supplementary Voltage Probe—B. Parzen. (PROC. I.R.E., vol. 37, pp. 1208–1211; October, 1949.) The impedometer consists of two oppositely connected directional couplers and a voltage probe in a short transmission line. An experimental model for frequencies between 50 and 500 Mc is described.

621.317.733 156
The Design and Construction of a Comparison Impedance Bridge for Frequencies of 40–270 Mc/s.—W. C. Weatherley. (*Proc. IEE* (London), vol. 96, pp. 429–432; September, 1949.) Two types of bridge are described, one for measuring "balanced" and the other for "unbalanced" impedances to earth. The frequency range is about ±20 per cent of the working mid-frequency. Measurements can be obtained without placing bulky apparatus near the impedance being measured. Changes in impedance over a definite frequency band are measured, rather than absolute impedance. Circuit details, suitable components, and calibration procedure are described. Accuracy is within ±2 per cent for resistance comparison and within ±5 per cent for capacitance comparison.

621.317.74:621.395.44 157
Transmission-Measuring Set for Low-Frequency Carrier Systems—J. Brundage and J. Zyda. (*Elec. Commun.*, vol. 26, pp. 204–208; September, 1949.) Description of the Type 902-A set, which includes a stabilized R-C oscillator covering the frequency range 300 to 40,000 cps. A 4-position rotary switch provides for sending, receiving, measuring, and calibration. Transmission power may be varied from −40 to +20 db relative to 1 mw, and receiver sensitivity is adequate. Self-calibration is arranged. The unit may be either of portable form or rack-mounted.

621.317.75 158
Harmonic Analyzer and Synthesizer—J. Lehmann. (*Electronics*, vol. 22, pp. 106–110; November, 1949.) Any wave form which can be represented by a Fourier sine or cosine series having up to 20 terms can be analyzed by this electromechanical instrument. Two dials are provided on which the amplitude and phase of each term can be set. For synthesis, the amplitudes and phases of the first 20 odd harmonics are set on the dials, and the response is plotted on a dynamometer-type recorder. For analysis, the amplitudes of the steps of a stepped-wave approximation to the wave form under investigation are set on the amplitude dials, and the frequency response is obtained as two plotted curves from twin recorders, in rectangular or polar coordinates. Circuit arrangements are discussed, and examples illustrating the operation of the instrument are included.

621.317.755:621.385.629.63/.64 159
Traveling-Wave Oscilloscope—Pierce. (*See* 264.)

621.317.761 160
Electronic Frequency Meter—(*Jour. Sci. Instr.*, vol. 26, p. 310; September, 1949.) Made by Airmec Laboratories Ltd., High Wycombe. Designed to measure frequencies 0–200 cps, 0–2 kc and 0–20 kc. Input signal can be between 0.1 and 20v; the wave may be of any shape having not more than 2 zeros per cycle. A circuit diagram is given showing the wave form at various stages. Pulses of recurrence frequency corresponding to the signal frequency are derived. The meter indicates the number of pulses received per second. Accuracy within 1 per cent is unaffected by mains voltage fluctuations not exceeding 10 per cent.

621.317.761 161
A New Frequency-Meter and Its Application—H. Hochrainer. (*Elektrotech. u. Maschinenb.*, vol. 66, pp. 288–291; October, 1949.) The current whose frequency is to be determined passes through a barretter and thence through a capacitor and a resistor respectively to opposite corners of a rectifier bridge. If the currents through the capacitor and resistor are equal, a moving-coil ammeter connected across the other diagonal will show no deflection. If, however, the frequency alters, the capacitive current will change and a deflection will be obtained on one side or the other of the instrument's center zero. Applications to frequency recording for power systems and to frequency control are considered.

621.317.763:621.392.26† 162
Wave-Guide Interferometers as Differential Wave-Meters—A. B. Pippard. (*Jour. Sci. Instr.*, vol. 26, pp. 296–298; September, 1949.) The advantages of nonresonant waveguide interferometers over resonant cavities as absolute differential wave meters are discussed, and two simple interferometers are considered in detail. An interferometer using a hybrid junction, and

operating similarly to Michelson's interferometer, can have extremely high sensitivity in detecting small changes of frequency: experimental results are discussed.

621.317.772 163
On a Valve-Type Phase Meter—J. Gilbert. (*Helv. Phys. Acta*, vol. 22, pp. 409–411; August 15, 1949. In French.) Description, with circuit diagram giving no component details, of a meter for reading directly the phase difference between two lf voltages of any wave form, and any amplitude between 20 v and 250 v. For frequencies in the range 20–500 cps the accuracy is of the order of $\pm 2°$. A phase meter based on similar principles has been described by Florman and Tait (1726 of 1949). See also 3483 of 1949 (Kretzmer).

621.317.79:621.396.615 164
Citizens Band Signal Generator—W. C. Hollis. (*Electronics*, vol. 22, pp. 77–79; November, 1949.) Construction details of a unit which contains a tunable concentric-line resonator in a Colpitts oscillator circuit using a subminiature tube. For earlier articles on Citizens Radio see 3482 of 1949 (Lurie) and back references.

621.317.79:621.396.822 165
A Video-Frequency Noise-Spectrum Analyzer—P. S. Jastram and G. P. McCouch. (Proc. I.R.E., vol. 37, pp. 1127–1133; October, 1949.) Requirements for noise-analyzer design are discussed. The heterodyne method of analysis is preferred to the tuned-circuit method because wide frequency coverage can be obtained with constant sensitivity and simple tuning arrangements. Design procedures can be worked out for obtaining adequate dynamic operating range and for suppression of errors. All circuits after the first modulator operate over fixed narrow frequency bands. A practical instrument for a frequency range of 50 kc to 10 Mc is described. The width of the analyzing pass band is 33 kc.

621.317.79:621.396.933 166
Distant Monitoring of Radio Transmissions for Naval and Aviation Services—J. Marique. (*HF* (Brussels), no. 3, pp. 71–80; 1949.) The subject is considered as affecting ship or aircraft security, navigation, and communications. Examples illustrate the results of measurements carried out at the special monitoring center established in Brussels since 1946. An outline is given of the methods adopted there for the measurement of frequency, the recording of signals in a given frequency band and analysis of the records.

621.317.794 167
A Ten Centimeter Broadband Bolometer Cavity—T. Miller. (*Tele-Tech*, vol. 8, pp. 28–31, 74; September, 1949.) A 10-ma fuse is used as the bolometer element of a cavity resonator for measuring powers from 100 μw up to several mw. Power reflection is <1 per cent for wavelengths between 9 and 10.5 cm. The cavity bridge circuit and methods of increasing bandwidth by means of a resonant diaphragm soldered into the waveguide are described.

621.319.4.089.6 168
The Calibration Curves for Kohlrausch Capacitors—G. Zickner. (*Elektrotechnik* (Berlin), vol. 2, pp. 317–320; November, 1948.) Description of a parallel-plate capacitor with variable plate distance, and discussion of the representation of its calibration curve in cartesian, double-logarithmic, or hyperbolic coordinates, and also of the best means for determining the error curve showing the departure from the exact hyperbolic law.

621.385.001.4:621.3.015.3 169
Surge Testing of High Vacuum Tubes—H. J. Dailey. (*Tele-Tech*, vol. 8, pp. 26–29, 60; October, 1949.) An experimental investigation with artificially generated flash-arcs. Results indicate that (a) adding series resistors in the anode circuit does not materially affect tube damage per flash-arc for low gas pressures, but does reduce the tendency towards arcing, (b) shielding the filament supports from the anode causes prompt failure, (c) internal sources of gas must be minimized, (d) initial gas pressure should be of the order 10^{-6} mm Hg to minimize flash-arc damage, (e) the filament exposed to flash-arc should be at maximum operating temperature, (f) the addition of a series resistor may increase flash-arc damage if the tubes are soft and remain soft.

621.396.621.001.4:621.3.015.3 170
Transient Phenomena in Radio Receivers—Carniol. (*See* 318.)

621.396.645.37.001.4:621.3.016.352:621.3.015.3 171
Examination of Amplifier Stability by Applying a Sudden D.C. Test Voltage—Carniol. (*See* 77.)

621.397.62.001.4:621.396.619 172
Measuring Modulation Depths of TV Signals—R. P. Burr. (*Tele-Tech*, vol. 8, pp. 32–35, 77; September, 1949.) Description of methods used and equipment required, especially for cases where only a small test signal is available. See also 3593 of 1947 (Buzalski).

621.317 173
Hochfrequenzmesstechnik [Book Review]—O. Zinke. Publishers: S. Hirzel, Leipzig, 2nd edn 1947, 253 pp. (*Elektrotechnik* (Berlin), vol. 2, p. x; November, 1948.) There are twelve main sections dealing with current, voltage, field-strength, and power measurements and with practically every type of measurement required in radio technique. "... an extraordinarily valuable help for every hf technician."

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.718.4:621.396.619.13 174
Measurement and Amplification of Small Displacements by Means of F.M.—P. Bricout and M. Boisvert. (*Rev. Gén. Élec.*, vol. 58, pp. 402–404; October, 1949.) An instrument is described in which small displacements of the plate of a specially constructed capacitor in an oscillatory circuit produce frequency variations at the mixer grid of a heptode in a Hartley circuit. This discriminator is due to Bradley (227 of 1947) and is relatively insensitive to amplitude variations; it operates linearly over a large part of the scale. Sensitivity is adjustable within wide limits; amplification is above 1,000 and relative error below 1 per cent. Application of the instrument in various pressure measurements are noted.

534.321.9.001.8:539.61.08 175
Evaluation of Adhesion by Ultrasonic Vibrations—S. Moses and R. K. Witt. (*Ind. Eng. Chem.*, vol. 41, pp. 2334–2338; October, 1949.) A direct, quantitative method for measuring the adhesion of organic coatings to both metallic and nonmetallic surfaces is described. Longitudinal ultrasonic vibrations are induced by an electrodynamic system in a metal cylinder, one end of which is threaded for attachment of the sample to be coated. The coating separates from the face of the sample when the force required to accelerate it exceeds the adhesion force at the interface. The accelerating force is determined by the frequency and amplitude of the vibration, and the mass and dimensions of the film.

535.247.4 176
Darkroom Light Meter—S. Becker. (*Electronics*, vol. 22, pp. 90–91; November, 1949.) A sensitive photocell unit samples the light falling upon an enlarging easel and gives the optimum exposure time directly in seconds. Complete construction details are included.

535.61-15:621.385.532 177
Cathode-Ray Presentation for Infrared Spectrometer—J. H. Jupe. (*Electronics*, vol. 22, pp. 189, 193; November, 1949.) Description of an instrument capable of scanning a band of width 3μ anywhere in the range $2–16\mu$, in a total time of about 15 seconds.

539.16.08 178
Electron Mobilities in Geiger-Müller Counters—H. den Hartog, F. A. Muller, and C. S. W. van Rooden. (*Physica's Grav.*, vol. 15, pp. 581–587; July, 1949. In English.)

539.16.08 179
Geiger Tube Quenching Circuit for a Negative High Voltage Supply—R. J. Watts. (*Rev. Sci. Instr.*, vol. 20, p. 699; September, 1949.) A modified Neher-Harper circuit. The Type CK569AX tube was selected for a portable design; other heater types of tube can be used.

539.16.08 180
Speed of Operation of Geiger-Müller Counters—H. den Hartog. (*Nucleonics*, vol. 5, pp. 33–47; September, 1949. Bibliography, p. 47.) Discussion of the delays caused by electron transit time, discharge development, discharge quenching, and dead-time. The work of various authors who have attempted to explain or measure some of the complex phenomena involved is reviewed. The resolution of G-M counters is compared with that of other counters.

539.16.08:621.386 181
Dead Time and Non-Linearity Characteristics of the Geiger-Counter X-Ray Spectrometer—L. Alexander, E. Kummer, and H. P. Klug. (*Jour. Appl. Phys.*, vol. 20, pp. 735–740; August, 1949.)

621.316.7.076.7.:543.712:676/677 182
Automatic Control of Moisture—R. V. Coles. (*Electronics*, vol. 22, pp. 82–86; November, 1949.) Electronic equipment makes corrections proportional to the difference from the desired moisture content as material passes through a capacitive measuring element. This type of control is specifically useful in the textile and paper industries.

621.317.083.7 183
Electrical Telemetering—N. A. Abbott. (*Radio and Televis. News, Radio-Electronic Eng. Supplement*, vol. 13, pp. 3–6, 29; October, 1949.) Discussion of primary detectors for the actual measurements, intermediate links for transmitting the information, and indicating or recording units. A block diagram of a typical complete telemetering system is included.

621.317.39:531.775 184
A Tachometer—A. E. Benfield. (*Rev. Sci. Instr.*, vol. 20, pp. 663–667; September, 1949.) Theory and operation of a simple instrument in which a dc electromotive force is generated by means of a magnet attached to the rotating member. A formula for the emf is given, and the effect of varying the position and orientation of the magnet is considered. The emf is directly proportional to the speed of rotation, but is rather small; possible ways of increasing it are briefly discussed.

621.317.39:[620.178.3+531.768 185
Modern Vibration Meters with Electrical Indication—H. Kohler. (*Elektrotechnik* (Berlin), vol. 3, pp. 301–310; October, 1949.) The physical principles of vibration and acceleration meters are considered and short descriptions are given of a wide variety of instruments, including capacitive, electrodynamic, and piezoelectric types. These fall into two groups: (a) with relatively low sensitivity for investigations dealing with machinery; and (b) with high sensitivity for use in seismology and in building and mining research. A fuller account is given of a piezoelectric acceleration meter with very high sensitivity; this has a longitudinally stressed Rochelle-salt crystal as indicating element. Results obtained with this meter are described.

621.365.5 186
Coupling Circuits for H.F. Heating—R. A. Whiteman. (*Radio and Televis. News, Radio-Electronic Eng. Supplement*, vol. 13, pp. 8–11, 27; October, 1949.) Design of such circuits for optimum power transfer and optimum loading of the power stage.

621.365.54† 187
[Vacuum] Induction Furnace for High-Temperature Ceramic Research—P. D. Johnson. (*Jour. Amer. Ceram. Soc.*, vol. 32, pp. 316–319; October, 1949.) An illustrated description. Power is supplied by a 9,600 cps generator driven by a 90-hp, 3-phase motor. Temperatures up to 2,300°C can be obtained by using Mo for the susceptor and the radiation shields.

621.365.54†:621.785.6 188
Multi-Purpose Inductance Hardening Units—(*Machinery* (London), vol. 75, pp. 413–415; September 22, 1949.)

621.365.55† 189
Temperature Behaviour in Electrically Heated Nonhomogeneous Bodies—W. F. Kussy. (*Elektrotechnik* (Berlin), vol. 3, pp. 323–326; October, 1949.) Calculations with particular application to porcelain coil-formers, for both short-period and continuous loading.

621.365.55†:621.317.3.011.5 190
Dielectric Loss with Changing Temperature—Whitehead and Rueggeberg. (*See* 137.)

621.38.001.8 191
Electronics in War and Peace—A. L. Whiteley. (*Engineering* (London), vol. 168, pp. 350–352; September 30, 1949.) Long summary of paper read before the British Association. The advantages of electronics for war purposes are high sensitivity and high speed of response. Industrial applications to welding and printing are discussed. The development of electronics in industry has been slower than expected because the installation of electronic devices often requires the re-design of machinery and the training of operating and maintenance staffs. The accuracy of electronic devices is often far greater than that required. Tube life must be lengthened. Electronic engineers must have a closer understanding of the processes they hope to improve.

621.38.001.8:629.13.052 192
An Electronic Pressure-Sensitive Transducer—G. Day. (*Jour. Sci. Instr.*, vol. 26, pp. 327–329; October, 1949.) The applied pressure varies the interelectrode spacing of a double-diode tube working under space-charge-limited conditions. The resultant changes in the two anode currents produce an out-of-balance voltage in a bridge circuit of which the diodes form two of the arms. The device is primarily designed for use as an aircraft altimeter.

621.383.001.8:535.61-15 193
Applying the Infrared Image Converter Tube—R. D. Washburne. (*Electronics*, vol. 22, pp. 150, 164; November, 1949.) Description of a British tube, Type CRI-143, in which conversion is achieved by means of a uniform field between the anode and the cathode. The image is not inverted, and the tube is much simpler than its American counterpart. For another account see 1710 of 1948 (Pratt).

621.383.001.8:536.587 194
Light-Sensitive Cells—(*Overseas Eng.*, vol. 23, pp. 128–129; November, 1949.) Applications in industry are discussed, with particular reference to automatic temperature control during welding and pre-heating operations.

621.384.611† 195
Betatron Injection into Synchrotrons—F. K. Goward. (*Proc Phys. Soc.*, vol. 62, pp. 617–631; October 1, 1949.) Factors influencing the proportion of electrons which may be trapped and accelerated in a transition from betatron to synchrotron operation are discussed; this transition is important in synchrotron design. Machines of various energies are studied; satisfactory agreement between theory and experimental results is obtained.

621.384.611.2† 196
A 30-Million Volt Synchrotron for Medical Use—(*Nature* (London), vol. 164, pp. 726–728; October 29, 1949.) A fuller description of the equipment noted in 3223 of 1949 (Martin).

621.384.611.2† 197
The Dynamics of a Synchrotron with Straight Sections—N. M. Blachman and E. D. Courant. (*Rev. Sci. Instr.*, vol. 20, pp. 596–601; August, 1949.)

621.384.612.1† 198
The Electron Cyclotron—W. J. Henderson and P. A. Redhead. (*Nucleonics*, vol. 5, pp. 60–67; October, 1949.) An experimental model, built by the National Research Council of Canada, is described. Magnetic field and applied rf are constant; the electron time lag per revolution due to the relativistic increase is made equal to an integral number of periods of the rf field. This type of accelerator may have distinct advantages over other types in the energy region of 10^7–10^8 ev.

621.385.833 199
Electron-Optical Shadow Method—(*Tech. Bull. Nat. Bur. Stand.*, vol. 33, pp. 106–108; September, 1949.) A technique based on extensive theoretical analysis and developed at the National Bureau of Standards by L. L. Marton; theoretical formulas have been derived by S. H. Lachenbruch. An electron lens system is used to produce a shadow image of a fine wire mesh placed in the path of an electron beam. From the distortion in the shadow pattern caused by deflection of the electrons as they pass through the field under investigation, accurate values of field strength can be computed. The method can be applied to fields of very small dimensions, such as the fringe fields from the small domains of spontaneous magnetization in ferromagnetic materials. The method is somewhat similar to the electron-optical schlieren method, but is much better adapted to precise determination of field intensity.

621.385.833 200
A New Type of Focusing in a Magnetic Lens Field—H. Slatis and K. Siegbahn. (*Phys. Rev.*, vol. 75, p. 1955; June 15, 1949.)

621.396.615:621-12 201
The Reciprocator—White and Lord. (*See* 66.)

PROPAGATION OF WAVES

535.42:538.56 202
On the Diffraction of an Electromagnetic Wave through a Plane Screen—J. W. Miles. (*Jour. Appl. Phys.*, vol. 20, pp. 760–771; August, 1949.) The screen is assumed infinitely thin and perfectly conducting. The diffraction problem is formulated, using generalized cylindrical coordinates, in terms of the generalized Fourier transform of the tangential electric field in the aperture. An integral equation for this transform is obtained. The power transferred through the aperture is calculated and expressed in a variational form of the Schwinger type. The significance and behavior of the aperture impedance is considered. Other formulations of the problem, including one involving Babinet's principle (1335 of 1947), are discussed. The variational formulation appears to provide a convenient link between the results for large wavelengths for which Rayleigh's static methods are valid, and those for small wavelengths where geometrical optics can be used. It appears to be superior to the Kirchhoff theory for any assumed aperture field. See also 1845 of 1948 (Levine and Schwinger).

535.42:621.396.81.029.6 203
Diffraction of High-Frequency Radio Waves Around the Earth—M. D. Rocco and J. B. Smyth. (Proc. I.R.E., vol. 37, pp. 1195–1203; October, 1949.) The results of height-gain measurements at 7 frequencies in the range 25 to 9,375 Mc for a nonoptical path in the Gila valley, Arizona, can be adequately explained by diffraction theory alone for the lower frequencies. Fields observed at the higher frequencies for low terminal heights were considerably stronger than those predicted by standard diffraction theory, even under meteorological conditions which were subnormal near the ground. These strong fields had rapid variations with time but no variation with height. The frequency distribution of the signal fluctuations agrees with a Rayleigh distribution. Interpretation of the data according to the waveguide theory of atmospheric propagation (507 and 2892 of 1947) is being investigated.

538.566 204
Transmission of Electric Waves through Wire Grids—W. Franz. (*Z. Angew. Phys.*, vol. 1, pp. 416–423; June, 1949.) The transmission of plane em waves through a system of parallel grids of identical grid-constant is calculated, assuming that the radius of the wire is small with respect to the wavelength and to the wire spacing. The radiation resistances occur as coefficients in a system of linear equations from which the excitation of the individual grids and also the transmission properties of the whole system can be determined. Comparison with the measurements of Esau, Ahrens, and Kebbel (2631 of 1939) shows satisfactory agreement.

538.566.2:535.13 205
The Propagation of Electromagnetic Waves through a Stratified Medium and Its WKB Approximation for Oblique Incidence—H. Bremmer. (*Physica,'s Grav.*, vol. 15, pp. 593–608; August, 1949.) The plane-wave solution of Maxwell's equations for a stratified medium is split into a series of terms, which have a simple physical meaning; the first of these terms constitutes the WKB approximation. By the introduction of a convenient Hertzian vector, the original vector problem is made scalar. The application of the saddle-point method to the individual terms of the series leads to simple geometric-optical approximations.

621.396.11 206
Velocity of Electromagnetic Waves—C. I. Aslaskon. (*Nature* (London), vol. 164, pp. 711–712; October 22, 1949.) The U. S. Air Forces used shoran extensively for measuring geodetic distances of from 67 to 367 miles. The results appeared to indicate that the velocity of em waves was 299,792 km. The method of observation was to determine the minimum of the sum of the distances between an aircraft and two ground stations as the aircraft flew across the line between the stations. The distances were corrected for the velocity of radio waves by a numerical integration along the ray path based on nearly simultaneous psychometric observations. See also 3488 of 1948 (Essen and Gordon-Smith) and 208 below.

621.396.11:551.510.535 207
Some Important Results of the Geometrical-Optical Properties of the Ionosphere—K. Rawer. (*Rev. Sci.* (Paris), vol. 86, pp. 481–485; May and June, 1948.) The results are presented of calculations of the paths of rays reflected from an ionosphere layer with a parabolic ionization distribution, the angle of projection of the rays having any value up to 90°. Curves are also given for the field strength and the muf factor as a function of distance from the transmitter. Values calculated for the radius of the zone of silence for different angles of projection are in good agreement with the values found from formula (12) of Appleton and Beynon's paper noted in 3290 of 1940.

621.396.11.029.64 208
The Measurement of the Velocity of Propagation of Centimetre Radio Waves as a Function of Height above the Earth: Part 2—The

Measurement of the Velocity of Propagation over a Path between Ground and Aircraft at 10,000, 20,000 and 30,000 ft.—F. E. Jones and E. C. Cornford. (*Proc. IEE* (London), vol. 96, pp. 447–452; September, 1949.) Continuation of 1441 of 1948 (Jones). Results of observations at two oboe ground stations are given. The most probable values for the mean velocity of propagation between ground and aircraft are 299,713 km, 299,733 km, and 299,750 km for aircraft heights of 10,000 feet, 20,000 feet, and 30,000 feet respectively.

621.396.11.029.66† 209

Propagation Characteristics of Tenth-Millimetre Waves—R. Franz. (*Radio Tech.* (Vienna), vol. 25, pp. 461–464 and 581–583; August and October, 1949.) Absorption, transmission, and selective-reflection properties of various solids for these waves are reviewed and the quartz-lens and residual-ray methods of isolating them are outlined. Absorption and reflection properties of gases and dipolar substances and resonance phenomena in rock salt and other materials are considered.

621.396.81:621.397.5 210

WSTV Field-Strength Report—Goldsmith. (*See* 250.)

621.396.812+538.566.3 211

A Survey of Ionospheric Cross-Modulation (Wave-Interaction or Luxembourg Effect)—L. G. H. Huxley and J. A. Ratcliffe. (*Proc. IEE* (London), vol. 96, pp. 433–440; September, 1949.) Existing theoretical and experimental knowledge is surveyed. The pioneer theory of Bailey and Martyn (1934 Abstracts, p. 199 and p. 606) is restated in a form which relates it more closely to standard ionosphere theory. The experimental results are summarized in form which enables the magnitude of the effect to be deduced approximately for any pair of stations. The way in which observations of cross-modulation can be used in ionosphere research is outlined. See also 1767 of 1949 (Cutolo and Ferrero).

621.396.812.029.62 212

Painless Prediction of Two-Meter Band Openings—W. F. Hoisington. (*QST*, vol. 33, pp. 22–25; October, 1949.) Several examples of long ranges for amateur communication are correlated with the corresponding weather maps. Long ranges are associated with areas of high barometric pressure. The trailing edges of such high-pressure areas are particularly important.

621.396.812.3.029.56 213

Tropospheric Effects in Short and Medium Radio Wave Propagation—W. J. G. Beynon. (*Nature* (London), vol 164, p. 711; October 22, 1949.) The results obtained by Heightman (2308 of 1949) are confirmed by measurements made in July, 1945, at Loth, Sutherland, on wavelengths between 60 and 300 m. Echo signals of appreciable amplitude appeared at ranges of 45 km and 65 km. They could be identified as reflections from mountains and were observed at all frequencies between 1 Mc and 5 Mc. Amplitude was often constant for several minutes or even hours, but sometimes varied by a factor of 4 or 5 to 1 in a few seconds. No well-defined variation in mean amplitude was found.

RECEPTION

621.396.621+621.397.62 214

Mass Production of Radio and Television Receivers—M. Alixant. (*Radio Tech. Dig.* (Franç), vol. 3, pp. 197–213; August, 1949.) Discussion of modern methods, from the prototype to the finished and tested article.

621.396.621 215

Features of French Broadcasting Receivers for the Season 1949–1950—J. Rousseau. (*TSF Pour Tous*, vol. 25, pp. 325–328; October, 1949.) A list is given, with details of tubes fitted in each receiver, type of circuit, and number of stages. Only a few models have a hf stage preceding the frequency changer.

621.396.621 216

A 400-Mc/s Receiver Front End Employing Subminiature Tubes and New Miniature Tuned Circuits—V. H. Aske. (*Sylvania Technologist*, vol. 2, pp. 2–5; October, 1949.) A new type of matched single-tuned input circuit is used, with plunger-type tuning. This is followed by a double-tuned rf stage and a pentode-type mixer working into a 30-Mc if stage. The if circuit is loaded with an impedance which simulates the input loading of the if tube. Design considerations and circuit and performance details are discussed. Gain could be increased by reducing the bandwidth of the rf tubes. Results show that tubes are available which can be used satisfactorily for the awkward frequencies between 400 and 1,000 Mc.

621.396.621:621.317.35 217

On the Energy-Spectrum of an Almost Periodic Succession of Pulses—G. G. Macfarlane. (Proc. I.R.E., vol. 37, pp. 1139–1143; October, 1949.) The energy/frequency spectrum is discussed for (a) regularly spaced pulses whose amplitudes have random oscillations about a mean value, and (b) identical pulses whose recurrence rate varies in a random manner about a mean value. Both spectra have two components, a line spectrum and a continuous spectrum. In case (a), the envelope of each component is proportional to the envelope of the spectrum of a single pulse and the spacing of the lines in the line spectrum equals the repetition frequency. In case (b), the envelopes of the two spectra are not the same as that of a single pulse, and the spacing of the lines equals the mean repetition frequency. Extension of the method of calculation to other cases is briefly discussed.

621.396.621.001.4:621.3.015.3 218

Transient Phenomena in Radio Receivers—B. Carniol. (*Tesla Tech. Rep.* (Prague), pp. 21–34; March, 1949.) For testing the af part of a receiver, square-wave oscillations with a relatively low fundamental frequency are used, so that any transients will terminate within less than half a period. Unstable freqencies well outside the af range usefully transmitted may affect receiver performance considerably. For the hf stages, a rectangularly modulated hf signal is used. Relations between frequency characteristics, phase characteristics, and transients are discussed. Transient phenomena occurring in tuned circuits and band-pass filters on the sudden application of a dc or hf voltage are considered in some detail and illustrated by oscillograms and phase characteristics. See also 77 above.

621.396.621.53 219

High Gain V.H.F. Converter—H. O'Heffernan. (*Short Wave Mag.*, vol. 5, pp. 720–724; February, 1948.) Design, construction, and component details for a 50 to 58 Mc unit of exceptional performance. The unit is arranged to have extremely short rf wiring. Each stage can be easily adjusted. Antenna coupling can be varied from the front panel. There is similar front-panel control of the first rf stage trimmer. Transformer or capacitive coupling is available between stages, and coils are easily accessible. The converter is designed to work into the common 1.6-Mc if channel of the main receiver, and to be switched in when required.

621.396.621.54 220

New Reception Principle for Superheterodynes—F. Tomek. (*Radio Tech.* (Vienna), vol. 25, pp. 584–586; October, 1949.) Description of a patented circuit termed the "summadyne" Instead of selecting the difference between the signal frequency and that of the local oscillator as the receiver if, the sum of the two frequencies is chosen. In order, therefore, to keep the if constant, the tuning of the local oscillator must be lowered as that of the input circuit is raised. This is effected by using identical components for the two circuits, with variable capacitors ganged in opposition. The advantages of such an arrangement are enumerated. A differential capacitor is used for tuning in two examples of receiver circuits, which both use double heterodyning and one of which has an amplifier stage for the first if. Intermediate frequencies of about 2,000 kc and 100 kc are used.

621.396.622 221

Signal-to-Noise Ratios of Linear Detectors—R. H. DeLano. (Proc. I.R.E., vol. 37, pp. 1120–1126; October, 1949.) A practical approach to the problem of obtaining the signal and noise spectra of the output of a linear detector, given the input signal and the input noise power spectrum. Graphical Fourier analyses are performed since the corresponding analytical expressions are cumbersome. The ratio of the input bandwidth to the center frequency is assumed small. Certain restrictions are assumed for the input signal: an am signal, for example, is regarded as the product of a slowly varying envelope and a carrier-frequency sine wave. A comparison is made with square-law detection for a few useful cases. The linear detector gives a higher output signal-to-noise ratio than the square-law detector for some types of signal.

621.396.622 222

An Improved Synchronous Detector—W. C. Michels and E. D. Redding. (*Rev. Sci. Instr.*, vol. 20, pp. 566–568; August, 1949.) The synchronous amplifier design of Michels and Curtis (43 of 1942) has been modified to minimize feed back and so to allow pre-amplification and a sensitivity of 2×10^8 mm/v. The instrument has an inductive input impedance of 2 MΩ and a bandwidth of 0.25 cps at 800 cps. Its merits are discussed.

621.396.622:537.312.62 223

Detec ion at Radio Frequencies by Superconductivity—J. V. Lebacqz, C. W. Clark, M. C. Williams, and D. H. Andrews. (Proc. I.R.E., vol. 37, pp. 1147–1152; October, 1949.) Detection by superconducting CbN was studied as a function of rf current, bias current, and temperature. A nonlinear resistance effect occurs in the transition region, with specially high values of dR/dI for currents <1 ma. This is believed to be due to the effect of the magnetic field of the current superconductivity (Silsbee hypothesis). The observed values of dR/dI explain to a certain extent the observed rectified potentials at 1 Mc, but the increasing rectification observed at higher frequencies is not yet explained. See also 2531 of 1949 (Lebacqz and Andrews).

621.396.622:621.396.619.13 224

Experimental Tube for F.M. Detection—L. J. Giacoletto. (*Electronics*, vol., 22, pp. 87–89; November, 1949.) A single-tube locked-in oscillator was used for FM detection by Bradley (227 of 1947) with a grid-controlled multigrid tube such as Type 6SA7. For optimum performance of this circuit, the current flowing beyond the second grid of the tube should not affect the oscillator circuit except through the feedback loop. In conventional multigrid tubes, the oscillator grid and the input-signal grid are not sufficiently isolated to permit the use of a high-impedance input circuit. To overcome this difficulty a tube combining beam deflection and grid control was developed. Audio output on an early model was low, but proposed design changes may increase the average output current at least fivefold. Performance characteristics of various FM detectors are compared.

621.396.622.7 225

Limiter Discriminator versus Ratio Detector—H. K. Milward and R. W. Hallows. (*Radio-Electronics*, vol. 21, pp. 20–22; November,

1949.) Results of tests carried out in England indicate that (a) although the difference in noise-rejection is not great, the discriminator with a single limiter has slightly better performance than the ratio detector, (b) the discriminator with 2 limiters is decidedly superior to the ratio detector, particularly when the noise level is high. Test apparatus and procedure are discussed. See also 2345 of 1948 (Maurice and Slaughter).

621.397.82 226
The Influence of U.H.F. Allocations on Receiver Design—Reid. (*See* 258.)

STATIONS AND COMMUNICATION SYSTEMS

621.391.5:621.395.623.66:791.44 227
Inductive Prompting System—B. H. Denney and R. J. Carr. (*Electronics*, vol. 22, pp. 66–69; November, 1949.) A system which allows the producer to communicate with actors in a film studio, and which does not interfere with the film sound system. A modulated magnetic field is detected and demodulated in receivers of the hearing-aid type. Receiver leads can be made photographically invisible or concealed under the actor's clothing. The receivers have neither tubes, nor batteries; crystal detectors are used. Each actor wears a coil of wire in which a secondary current flows; the 100-kc transmitter, of which circuit details are given, is connected to a single-turn loop which surrounds the set area and induces a strong rf field at all points within the area.

621.396.1:621.396.931 228
Radio Communications Services: Parts 2–4—(*FM-TV*, vol. 9, pp. 18–21, 45, 21–24, 30, and 21–23; July, September, and October, 1949.) Part 1: 2907 of 1949.

621.396.619 229
Neglected Outphasing System of Modulation—W. H. Hartman. (*CQ*, vol. 5, pp. 18–26, 68; October, 1949.) Theory, design, and constructional details of a practical transmitter using this form of modulation, which combines all the advantages of high-level modulation of a class-C amplifier with the economies of low-level systems. See also 85 of 1936 (Chireix).

621.396.619.16 230
A Method of Asymmetrical Pulse Duration Modulation—R. J. Watts. (*Rev. Sci. Instr.*, vol. 20, pp. 622–623; August, 1949.)

621.396.65+621.396.931/.932 231
Review of the Applications of V.H.F. Radio Communications—E. W. Northrop. (*GEC Jour.*, vol. 16, pp. 184–196; October, 1949.) Discussion of typical equipment both for mobile and for point-to-point communication choice of headquarters site, antenna height, optimum power and frequency, etc

621.396.712 232
The "New Look" at KTBS—(*Broadcast News*, no. 56, pp. 40–45; September, 1949.) General description of facilities available, with photographs of studios, control room, etc.

621.396.931 233
The Planning of "Business-Radio" Services at Very High Frequencies—(*Proc. IEE* (London), vol. 96, pp. 381–382; September, 1949.) Report on IEE Radio Section discussion meeting. Two main types of service are considered: (a) an extension of the ordinary telephone service to mobile units, and (b) an exclusive service for the various mobile units within one organization, such as a taxi company.

621.396.97 234
25 Years Broadcasting in Austria—W. Füchsel. (*Radio Tech.* (Vienna), vol. 25, pp. 576–580; October, 1949.) A review of developments at the RAVAG station, Vienna. The 120 kw transmitter installed in 1932 on the Bisamberg, the 100-kw Graz-Dohl transmitter and low-power transmitters in Karnten and Steiermark are mentioned. Subscribers now number over 1,200,000.

SUBSIDIARY APPARATUS

621-526 235
Servomechanisms and Modern Physics—R. Moch. (*Radio Tech. Dig.* (Franç), vol. 3, pp. 133–145 and 235–249, 252; June and August, 1949.) The construction and operation of servomechanisms is studied and also the electronic or electromechanical calculating elements which can be incorporated in such devices, including differentiators, integrators, and apparatus for other types of mathematical operations. Various examples of the application of servomechanisms are briefly described, these include process control, stabilization (as in the "atomic clock" of the National Bureau of Standards), and calculating machines.

621-526 236
Design Equations for Servomechanisms—B. Parzen. (*Elec. Commun.*, vol. 26, pp. 249–256; September, 1949.) Reprinted from the book noted in 269 below. Fundamental quantities, and relations between them, are derived for linear lumped constant servomechanisms; positioning system using electronic and electromechanical devices are considered in more detail. General theory and its application to typical positioning systems are discussed. Performance and stability criteria are briefly mentioned. See also 232 of 1949 (James, Nichols, and Phillips).

621-526 237
Magnetic Fluid Clutch in Servo Applications—G. R. Nelson. (*Electronics*, vol. 22, pp. 100–103; November, 1949.) Report on experience obtained with various iron-disk rotor designs running in a mixture of oil and powdered iron which solidifies when a magnetic field is applied. Such clutch units are useful in servomechanisms at natural frequencies below 30 cps.

621.3.076.7:621.3.016.1 238
Torque and Speed Regulation with the Electronic Amplidyne—J. L. Dutcher. (*Elec. Mfg.*, vol. 42, pp. 84–89, 158; July, 1948.) Principles of the amplidyne, which differs from a conventional generator only in having an extra pair of short-circuited brushes, are discussed. Circuit diagrams showing its use for various applications are included.

621.3.077.2/.3 239
The Amplidyne Generator—Its Performance and Design—M. S. Hoffenberg. (*Trans. S. Afr. Inst. Elec. Eng.*, vol. 40, pp. 175–188; August, 1949. Discussion, pp. 188–191.) Discussion of (a) the desirability of high amplification and low time constant in control appratus, (b) the cross-field principle of operation of the amplidyne, its steady-state operation on dc, and the effect of varying the compensation of the output stage, (c) the transient and steady-state response to ac signals of both perfectly and imperfectly compensated meachines, (d) the effects of rocking the brushes and chording the armature windings, (e) two types of stator construction, (f) the use of an output coefficient as an aid to design, (g) the effect of field form factor on the average flux density, and of quadrature current on specific electric loading, (h) variation of power-amplification ratio with number of cycles, (i) the magnetic circuit and commutation, and (j) the calculation of leakage fluxes and stage inductances.

621.314.6:621.315.59 240
Semiconductor Rectifiers—(*Elec. Eng.*, vol. 68, pp. 865–872; October, 1949.) Long summary, compiled by S. J. Angello, of 4 papers read at an AIEE symposium on "Electrical Properties of Semiconductors and the Transistor," namely: "Theory of rectification," by F. Seitz; "Boundary layers in rectifiers," by H. Y. Fan; "Noise in semiconducting contacts," by P. H. Miller, Jr.· "A comparison between the Schottky rectifier theory and measurements upon cuprous oxide cells," by S. J. Angello.

621.314.63 241
High Inverse Voltage Germanium Rectifiers—S. Benzer. (*Jour. Appl. Phys.*, vol. 20, pp. 804–815; August, 1949.) Current/voltage characteristics were determined for various point-contact Ge rectifiers. Inverse voltages of several hundred volts were observed; a reproducible negative differential-resistance region occurs in the inverse characteristic. The metal used for the point contact has little effect; the effects of impurities, surface treatment, temperature, and the force of contact are discussed. Contact between two Ge crystals is also considered.

621.314.63 242
The Characteristics and Applications of Metal Rectifiers—P. A. Goodyear. (*Trans. S. Afr. Inst. Elec. Eng.*, vol. 40, pp. 147–172; July, 1949.) The characteristics of Se rectifiers are briefly described, and applications are considered fully. The relative merits of Se and Cu_2O rectifiers are discussed.

621.314.67 243
Thoriated Tungsten Filaments in Rectifiers—Z. J. Atlee. (*Elec. Eng.*, vol. 68, p. 863; October, 1949.) Summary only. A thoriated tungsten filament can have an emission efficiency of 30 to 40 ma/w at an operating temperature of 1,800 to 2,200° k with a life of 2,000 hours; it is thus superior to a pure tungsten filament. An account is given of the development of a very small 25-w rectifier for use in an X-ray application requiring a maximum of 110 kv and 250 ma from a full-wave bridge rectifier with 4 rectifier tubes. Loss of emission by positive ion bombardment is prevented by a new method of applying adequate getter. A new seasoning process prevents the deposit of thorium on the anode and glass.

621.316.722.1 244
Operation of Voltage-Stabilizing Elements with Current-Stabilized Supplies—J. J. Gilvarry and D. F. Rutland. (*Rev. Sci. Instr.*, vol. 20, pp. 633–637; September, 1949.) The performance parameters are determined. The range-regulation factor Q is defined and suggested as a figure of merit. Advantages of a current-stabilized power supply for voltage stabilization under varying load conditions are emphasized. Stability criteria for the case when the stabilizing element has a negative-resistance characteristic are given, and also a graphical analysis for the case of nonlinear elements. The special case of voltage-reference tubes is discussed, with experimental results.

621.396.68 245
Variable Frequency Power Supply—E. B. Steinberg. (*Elec. Mfg.*, vol. 43, pp. 100–102; May, 1949.) Design details of a single-phase parallel inverter, using hydrogen thyratrons Type 5C22, which has an output of several kilowatts. The output frequency can be adjusted between 60 and 5,000 cps. Performance data are discussed.

621.316.7.004.5:621.38 246
Maintenance Manual of Electronic Control [Book Review]—R. E. Miller (Ed.). Publishers: McGraw-Hill, New York, 1949, 304 pp. $4.50. (*Electronics*, vol. 22, pp. 235–236; November, 1949.) Based on a series of articles published in *Electrical Construction and Maintenance* on the installation and service of electronic equipment. "The book should be a valuable aid to any engineer, maintenance man, or technician who is involved in any way with timing relays, time-delay relays, photoelectric relays, electronic motor control, welding con-

trol, furnace temperature control. and mercury-arc rectifiers."

TELEVISION AND PHOTO-TELEGRAPHY

621.397.331 2 **247**
The Image Isocon—An Experimental Television Pickup Tube Based on the Scattering of Low Velocity Electrons—P. K. Weimer. (*RCA Rev.*, vol. 10, pp. 366–386; September, 1949.) A description of a new method of generating the video signal by introducing additional helical motion into the primary beam and collecting the electrons not reflected specularly to give an output signal of polarity opposite to that of the maximum current in the light. This results in (a) an improved signal-to-noise ratio in the darker parts of the picture, and (b) freedom from spurious signals caused by multiplier dynode spots, an advantage which offsets the slightly superior resolution of the image orthicon. Closer tolerances in component design are required. The task of the operating crew is more exacting and the time lag is more objectionable than in the case of the image orthicon. Applications of the techniques used in the isocon are suggested for color television and for image-storage tubes.

621.397.5 **248**
How RCA's Color TV Works—E. W. Engstrom. (*FM-TV*, vol. 9, pp. 11–13, 30; October, 1949.) The principles of the system were discussed in 3297 of 1947 and 572 of 1948. Improvements have been added which make possible the transmission of a high-definition color picture in a 6-Mc channel. Present receivers need no modification to receive these color transmissions in monochrome. A block diagram of the broadcasting station is included and discussed. See also *Electronics*, vol. 22 pp. 122, 189; November, 1949.

621.397.5:535.88:791.45 **249**
Theater Television Today—J. E. McCoy and H. P. Warner. (*Jour. Soc. Mot. Pic. Eng.*, vol. 53, pp. 321–350; October, 1949.)

621.397.5:621.396.81 **250**
WDTV Field-Strength Report—T. T. Goldsmith, Jr. (*FM-TV*, vol. 9, pp. 15–18, 30; September, 1949.) Field-strength measurements for a hilly path radiating from Pittsburgh are tabulated, shown graphically, and compared with those expected from the results given in the Ad Hoc Committee's report (3524 of 1949).

621.397.5(437) **251**
Television in Czechoslovakia—J. Havelka. (*Tesla Tech. Rep.* (Prague), pp. 2–3; March, 1949.) Post-war work has hitherto only been possible in laboratories. Only locally produced equipment is used; this includes a kinescope with a flat square front. 625 lines, 25 frames per second, and negative modulation are used; picture and sound transmission are separate.

621.397.62+621.396.621 **252**
Mass Production of Radio and Television Receivers—Alixant. (*See* 214.)

621.397.62 **253**
New Television Receiver without Transformers. Design with Interchangeable Units—R. Aschen. (*TSF Pour Tous.* vol. 25, pp. 329–330; October, 1949.) Continuation of 2658 of 1949. Full details are given of the sound receiver, which uses two Type-UF41 pentodes as hf amplifier and grid detector respectively, and a Type-UL41 pentode for the output stage. Sensitivity is of the order of 200 to 300 μv for 50 mw in the loudspeaker.

621.397.62:621.385 **254**
Radio-Frequency Performance of some Receiving Valves in Television Circuits—R. M. Cohen. (*Radiotronics*, no. 138, pp. 58–63; July and August, 1949.) Reprint. See 2971 of 1948.

621.397.645 **255**
Stagger-Tuned Television Amplifiers—A. Easton. (*Radio Tech. Dig.* (Franç), vol. 3, pp. 147–152; June, 1949.) French version of 1344 of June.

621.397.7 **256**
WBAL-TV Channe 11, Baltimore—W. C. Bareham. (*Broadcast News*, no. 56, pp. 72–80; September, 1949.) General description of facilities available. with photographs of studios, control room, etc. and map showing coverage.

621.397.8 **257**
TV Reception below Line of Sight—R. B. McGregor (*Electronics*, vol. 22, pp. 72–76; November, 1949.) Signals were received at a point 92 miles from the transmitter and 2,000 feet below the line of sight with the aid of an 18-element antenna array, a cascode preamplifier (3061 of 1948) at the antenna and special if amplifier stages and sweep circuits. Signals were received whenever the transmitter was operating. About 50 per cent of the time, the picture was satisfactory. Sometimes it remained consistently good all the evening. Sometimes it was good by day and poor by night; more often vice versa. Some frequency-selective reception was noted. No correlation between reception and weather could be found.

621.397.82 **258**
The Influence of U.H.F. Allocations on Receiver Design—J. D. Reid. (Proc. I.R.E., vol. 37, pp. 1179–1181; October, 1949.) Added protection of receivers from local oscillator radiation can be obtained by alternate channel assignment and receiver if standardization. Additional protection from image interference can be given by having the picture-signal station separated by $\sqrt{2}$ times the distance between stations on adjacent channels, and co-channel stations separated by double this distance. 41.25 Mc is suggested as the optimum if for joint vhf/uhf usage.

TRANSMISSION

621.396.61:621.396.662 **259**
A New Type of V.H.F. Tank Design—Parker. (*See* 79.)

621.396.619.23 **260**
Rectifier Modulators with Frequency-Selective Terminations—D. G. Tucker. (*Proc. IEE* (London), vol. 96, pp. 422–428; September, 1949.) A simple method is given for determining the performance of such a modulator provided that the terminating impedance is resistive, zero, or infinite at all significant frequencies. Cases likely to be useful or unavoidable in practice are worked out for the ring, Cowan (shunt-type), and series modulators. Optimum terminating resistances and minimum insertion losses are tabulated in terms of the rectifier ratio and geometric mean resistance.

VACUUM TUBES AND THERMIONICS

621.385:621.396.645 **261**
Operation of Output Valves in High-Power Public-Address Amplifiers—Bezladnov. (*See* 75.)

621.385:621.397.62 **262**
Radio-Frequency Performance of some Receiving Valves in Television Circuits—R. M. Cohen. (*Radiotronics*, no. 138, pp. 58–63; July and August, 1949.) Reprint. See 2971 of 1948.

621.385.001.4:621.3.015.3 **263**
Surge Testing of High Vacuum Tubes—Dailey. (*See* 169.)

621.385.029.63/.64:621.317.755 **264**
Traveling-Wave Oscilloscope—J. R. Pierce. (*Electronics*, vol. 22, pp. 97–99; November, 1949.) The 1,000-v oscilloscope here described was developed specially for laboratory examination of short recurrent pulses. Response is almost flat from 0 to 500 Mc. Input impedance is 75Ω. A peak-to-peak signal of 0.37 v gives a pattern 10 trace-widths high, which is viewed through a microscope. Good vertical deflection sensitivity is obtained by means of a traveling-wave deflection system, without transit-time bandwidth limitations. The essentials of the oscilloscope are illustrated and discussed.

621.385.032.21:621.317.39:536.5 **265**
On a New Method of Measuring the Temperature of a Thermionic Cathode—P. Gandin and R. Champeix. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 545–547; September 12, 1949.) By differentiation of the expression for the anode current I in terms of the saturation current, negative voltage of the anode, and cathode temperature T, the following formula is derived: $T=11,600\,\rho\, I$, where ρ is the differential resistance of the cathode/anode space in ohms, I is in amperes and T in absolute degrees. A simple measurement circuit is described; this makes use of a transformer whose secondary provides two equal voltages of opposite phase. These voltages should be less than 0.01 v. One voltage is applied to cathode and anode, a variable dc voltage and a galvanometer being included in the cathode lead; the other voltage feeds a standard resistor, one terminal of which is connected to the anode. A null indicator is connected in the lead common to the two branches. Errors in determining T should not exceed 2 per cent. Results obtained by this method will be given in a later paper

621.385.032.212:681.142 **266**
Polycathode Glow Tube for Counters and Calculators—J. J. Lamb and J. A. Brustman. (*Electronics*, vol. 22, pp. 92–96; November, 1949.) A cold-cathode neon-filled discharge tube whose basket-shaped anode has 30 narrow slots. Cathode fingers, in staggered sets of 10, are located so as to line up with the anode segments, the gap being 0.020 inch. The tenth finger of one set is separate from the remainder and is used for numerical carrying. The initial potential difference between anode and cathodes is such that a discharge takes place between only one of the cathode fingers and the anode when a potential exceeding the breakdown voltage is applied. The residual ionization around this finger favors the formation of the next discharge at the adjacent finger of another cathode ring when that ring is energized. The tube is capable of operation at rates up to 10^5 per second and can be adapted to decade counting circuits with rates exceeding 16,000 per second.

621.385.832:535.371.07:621.396.9 **267**
Radar Screens—B. de la Pinsonie. (*Bull. Soc. Franç. Élec.*, vol. 9, pp. 532–542; October, 1949.) The mechanism of fluorescence and phosphorescence is discussed, the characteristics of screens specially adapted for radar presentation are considered and suitable methods for measuring the various characteristics are described, with experimental results, both for ordinary radar screens and for the skiatron type.

621.396.645 **268**
Amplification by Direct Electronic Interaction in Valves without Circuits—P. Guénard, R. Berterottière, and O. Doehler. (*Bull. Soc. Franç. Élec.*, vol. 9, pp. 543–549; October, 1949.) See 2977 of 1949.

MISCELLANEOUS

621.396 **269**
Reference Data for Radio Engineers [Book Review]—Publishers: Federal Telephone and Radio Corporation, New York, 3rd edn. 1949, 672 pp., $3.75. (*Elec. Commun.*, vol. 26, p. 242; September, 1949.) Material in the second edition (See 1301 of 1947) has been expanded and new material added, doubling the size. See also 236 above.